SOLID-STATE CONVERSION OF A TUBE-TYPE COMMUNICATIONS RECEIVER

James Michael Steussy

NAVAL POSTGRADUATE SCHOOL
Monterey, California

THESIS

Solid-State Conversion of a Tube-Type
Communications Receiver

by

James Michael Steussy

Thesis Advisor: Charles H. Rothauge

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

REPORT DOCUMENTATION PAGE

READ INSTRUCTIONS BEFORE COMPLETING FORM

1. REPORT NUMBER	2. GOVT ACCESSION NO.	3. RECIPIENT'S CATALOG NUMBER
4. TITLE (and Subtitle) Solid-State Conversion of a Tube-Type Communications Receiver		5. TYPE OF REPORT & PERIOD COVERED Master's Thesis; September 1974
		6. PERFORMING ORG. REPORT NUMBER
7. AUTHOR(s) James Michael Steussy		8. CONTRACT OR GRANT NUMBER(s)
9. PERFORMING ORGANIZATION NAME AND ADDRESS Naval Postgraduate School Monterey, California 93940		10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS
11. CONTROLLING OFFICE NAME AND ADDRESS Naval Postgraduate School Monterey, California 93940		12. REPORT DATE September 1974
		13. NUMBER OF PAGES
14. MONITORING AGENCY NAME & ADDRESS(if different from Controlling Office) Naval Postgraduate School Monterey, California 93940		15. SECURITY CLASS. (of this report) Unclassified
		15a. DECLASSIFICATION/DOWNGRADING SCHEDULE

16. DISTRIBUTION STATEMENT (of this Report)

Approved for public release; distribution unlimited

17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)

Approved for public release; distribution unlimited

18. SUPPLEMENTARY NOTES

19. KEY WORDS (Continue on reverse side if necessary and identify by block number)

solid-state receiver conversion

20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

A Hammarlund SUPER-PRO SP-210 communications receiver of approximately 1940 design was rebuilt using modern solid-state components and associated circuitry in replacement of the original eighteen vacuum tubes. Replacement was effected in a stage-for-stage manner so as to utilize, wherever practicable, the original system of operation and particularly the existing tuned circuits for which the receiver was famous. The project goals were to gain experience in the applications of classroom theory to practical electronics, to observe the problems associated with both vacuum tube and solid-state circuitry, and to

SECURITY CLASSIFICATION OF THIS PAGE (When Data Entered)

provide a starting point for continued research in electronics.

Solid-State Conversion

of a Tube-Type

Communications Receiver

by

James Michael Steussy
Lieutenant, United States Navy
B. S., United States Naval Academy, 1968

Submitted in partial fulfillment of the
requirements for the degree of

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

from the

NAVAL POSTGRADUATE SCHOOL

September 1974

ABSTRACT

A Hammarlund SUPER-PRO SP-210 communications receiver of approximately 1940 design was rebuilt using modern solid-state components and associated circuitry in replacement of the original eighteen vacuum tubes. Replacement was effected in a stage-for-stage manner so as to utilize, wherever practicable, the original system of operation and particularly the existing tuned circuits for which the receiver was famous. The project goals were to gain experience in the applications of classroom theory to practical electronics, to observe the problems associated with both vacuum tube and solid-state circuitry, and to provide a starting point for continued research in electronics.

TABLE OF CONTENTS

LIST OF TABLES

LIST OF ILLUSTRATIONS

I. PROJECT DESCRIPTION

A. PROJECT BACKGROUND

1. History of the SUPER-PRO

The Hammarlund Manufacturing Company was a principal competitor in the production of radio receivers for military, commercial, and amateur use during the 1930's and 1940's. In the late thirties, Hammarlund began advertising the first of a series of communications receivers which were given the trade-name "SUPER-PRO." The receiver described herein was an SP-210 model SUPER-PRO built for consumer use. It was a second-generation SUPER-PRO and was first advertised in 1941 for a list price of $279.95. During the World War II period, an almost identical receiver, the BC-779, was produced for the U. S. Armed Forces as part of the SCR-244A, SCR-704, and AN/FRR-4 radio communications sets. The specifications of this receiver are listed in Appendix A. Variants of the receiver covering other frequency bands were also produced for the military. After the war's end, these fine receivers became prizes in the surplus markets of the world and there are probably thousands of war-vintage SUPER-PROs still in service. The author was fortunate to acquire two such receivers (one SP-210 and one BC-779A) and a matching power supply.

The possibilities of an interesting and educational project of SUPER-PRO modernization were illustrated in detail by Knietel [Ref. 1]. The conversion offered therein

used miniature vacuum tubes and featured a built-in solid-state power supply. The results suggested that a completely solid-state conversion of the SUPER-PRO would be a rewarding effort in terms of practical experience in electronics and appreciation of the capabilities, advantages, and limitations of both vacuum tube and semiconductor active devices. On this premise, this project was undertaken.

2. Project Goals

The purpose of this project was to develop an understanding based on experience of practical electronics. The transition "from blackboard to breadboard" is often a difficult one and an engineering appreciation for this situation was the primary objective of the project. Certainly an easier project would have been the design and construction "from scratch" of a solid-state receiver which would have been smaller, lighter, and probably better in terms of performance than a converted SUPER-PRO. However, the latter offered an opportunity to study vacuum tube circuits as well as state-of-the-art semiconductor circuits, clearly a valuable experience in view of the fact that much vacuum tube equipment remains operational in the U. S. Navy of 1974. Finally, the converted receiver would provide the foundation of a continuing project of experiments and improvements which would assist the author to keep up to date with the continuously evolving world of integrated electronics.

B. THEORY OF OPERATION

1. Principles of Superheterodyne Operation

The SUPER-PRO is an excellent example of the classic superheterodyne receiver design. The heterodyne principle, introduced by Fessenden of the University of Pittsburgh in 1913, is essentially the mixing of two signals of different frequencies to produce a signal at a desired third frequency. This principle was used by E. H. Armstrong in France during World War I when he developed the so-called "supersonic heterodyne" receiver which used a high frequency oscillator to convert received signals to a lower supersonic frequency, now known as the intermediate frequency. This type of receiver is now called the superheterodyne receiver and a simple block diagram is shown in Fig. 1. The superheterodyne receiver is a vast improvement over previous types because its selectivity and sensitivity are essentially independent of signal frequency. Today, almost all radio receivers of every kind, including radar, electronic warfare, color television, and even the cheapest pocket-sized broadcast types, utilize the superheterodyne principle of operation.

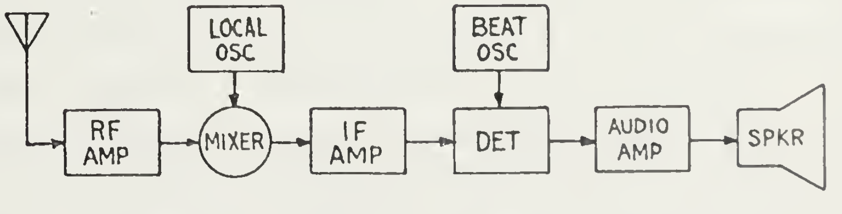

FIG. 1: BASIC SUPERHETERODYNE BLOCK DIAGRAM

As shown in Fig. 1, the first stage of the basic superheterodyne is the radio frequency (RF) amplifier. This stage is optional and is not included in cheaper AM and FM broadcast radios as well as many older microwave receivers. The primary purposes of the RF amplifier are establishment of a low noise figure and rejection of unwanted image frequency signals. The RF amplifier is usually frequency-tuned simultaneously with the mixer and local oscillator stages to provide single knob tuning. The local oscillator generates a signal which is greater (or less) than the desired RF signal by an amount exactly equal to the intermediate frequency. For proper heterodyne operation, the local oscillator signal voltage is at least 20 dB over, or 10 times as large as the RF signal. Both the local oscillator signal and the RF signal are inputs to the mixer. The outputs of a heterodyne mixer consist of signals at four separate frequencies. Two of these are the original components and the others are at the sum and difference frequencies. Only one of these signals, usually the difference frequency signal, is at the intermediate frequency. Since this difference frequency remains unchanged regardless of the frequency to which the receiver is tuned, fixed-frequency tuned amplifiers may be used following the mixer stage. The intermediate frequency (IF) amplifier provides the most important gain or amplification in the receiver and, to a large extent, the selectivity as well. The detector stage recovers amplitude modulation originally impressed on the signal in the transmitter and in which the signal information is contained. A separate oscillator, called the beat frequency oscillator (BFO,) is used to recover CW signals by a second application of the heterodyne process. The recovered modulation is then amplified again and is ready for use. In Fig. 1, the recovered information is a signal at audio frequency, which is used to drive a speaker.

The receiver depicted in Fig. 1 is a superheterodyne in perhaps its simplest form. Other circuits, such as automatic gain or frequency controls, noise reduction circuits, limiters, additional mixers and IF amplifier stages, filters, etc., are frequently included. This adds to the cost and complexity of the receiver, but permits the superheterodyne to be almost universal in application and usefulness.

2. Original SUPER-PRO Operation

The unmodified SUPER-PRO SP-210 receiver was a rather advanced device for its time. The block diagram, shown in Fig. 2, is considerably more sophisticated than the basic unit of Fig. 1. Two tuned RF amplifiers preceed the mixer and local (high frequency) oscillator. All four stages were simultaneously tuned by a pair of large four-gang air variable capacitors, which were the MAIN TUNING and BANDSPREAD controls on the front panel. These stages comprised the "front-end" of the SUPER-PRO and provided an intermediate frequency signal of 465 kHz. to the IF amplifier stages via the crystal filter. The crystal filter was a frequency selective device built around a piezoelectric crystal tuned to the intermediate frequency. Its operation is described in detail below. Three IF amplifier stages made up the IF "strip." Output of the second IF amplifier was sampled and amplified by the automatic volume control (AVC) amplifier, converted to a negative direct voltage in the AVC rectifier, and returned to the control grids of the tubes in the first and second RF and IF amplifiers. This direct voltage varied the bias on the tubes in such a manner as to make the amount of overall gain in these stages inversely proportional to signal strength. In this way, weak signals were amplified more

than stronger ones. The BFO signal was injected to the third IF amplifier plate and enabled the detector to demodulate CW signals. The noise limiter stage clipped high-amplitude noise pulses at the peak value of the strongest received signal and reduced their effect on the listener. Three stages of audio amplification, the last a powerful push-pull Class B power amplifier, followed the detector and provided more than enough volume for any listener. The power supply unit was built on a separate chassis and provided the positive plate and screen voltages, the negative grid bias voltage, and 6.3 volt filament and lamp alternating voltage.

A total of eighteen vacuum tubes, including the two rectifier tubes, were used in the SUPER-PRO SP-210 and BC-779.

3. Modified SUPER-PRO Operation

One of design criteria in the conversion of the SUPER-PRO receiver was to retain the character of the original design by following, as closely as possible, the same signal processing scheme and to use the front end and IF strip tuned circuits. The front end, therefore, was little modified beyond the replacement of the active circuit elements and associated components. Tuned circuits, tuning and bandswitching mechanisms were retained. The most significant front end modification was the addition of a crystal-controlled frequency marker generator at the antenna input. The crystal filter was retained largely unchanged and three stages of IF ampliciation were used. A noise blanking circuit, which samples the mixer output, cuts off the signal during high amplitude noise spikes. The modulated IF signal is then coupled to three detector stages. AM and FM signals are demodulated by the AM and FM

detectors, respectively, as shown in Fig. 3. A squelch circuit was included in the FM signal line to provide a squelched listening capability for that mode. CW and SSB signals are demodulated in the product detector stage with the aid of a BFO signal provided by one of three oscillators in the BFO module. Two oscillators are crystal-controlled for fixed detection of upper and lower sideband signals. The third oscillator is variable as in the original SUPER-PRO to provide greater flexibility in reception of CW signals than is possible with the crystal-controlled oscillators. Output of the three detectors is selected by the front panel MODE switch and is coupled to the audio module. A sample of this signal is also amplified and rectified by the automatic gain control (AGC) amplifier for controlling the gain of the second RF amplifier and all three IF amplifiers in much the same manner as the original AVC stages performed in the unmodified SUPER-PRO. The DC AGC signal is also fed to the S-meter amplifier, which drives the signal strength meter on the front panel. The audio module is composed of an audio frequency filter, a compressor (audio AGC amplifier,) a preamplifier, and a power amplifier, which is capable of driving a speaker. The power supply, mounted on the receiver chassis, provides regulated positive and negative voltages to all circuits.

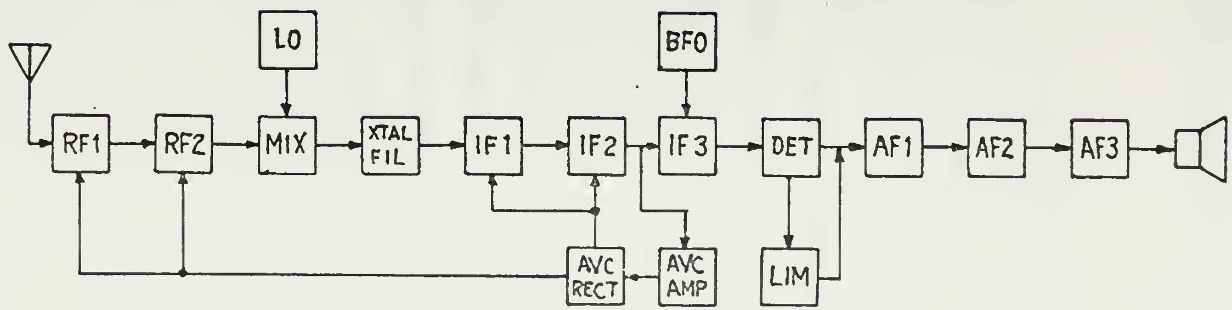

FIG. 2: ORIGINAL SUPER-PRO BLOCK DIAGRAM

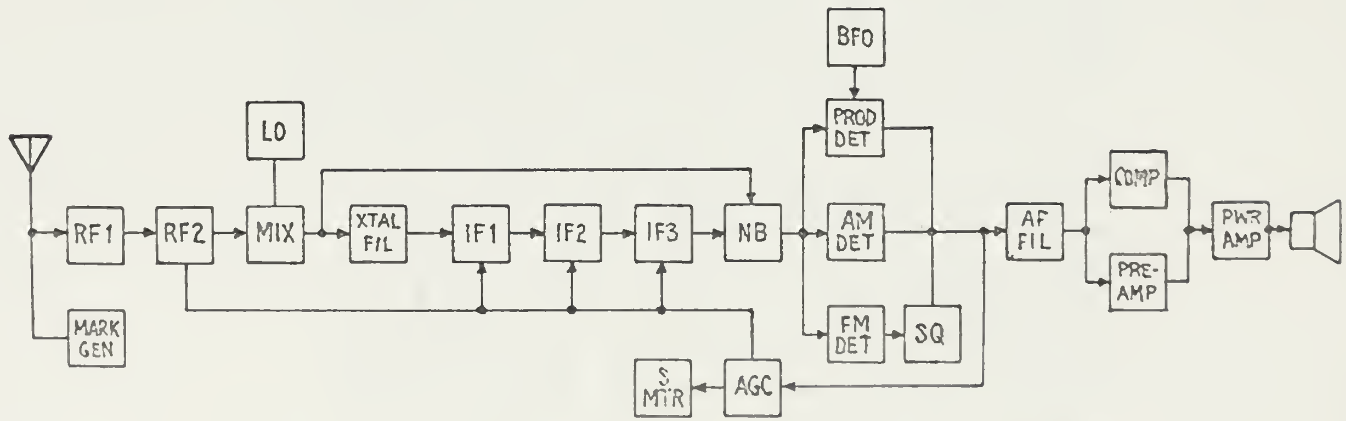

FIG. 3: MODIFIED SUPER-PRO BLOCK DIAGRAM

A total of twenty integrated circuits and twenty-seven transistors are used in this modification, including those employed in the power supply circuits.

C. CONVERSION DETAILS

As previously mentioned, two SUPER-PRO receivers were available. The BC-779A was rusty and moldy externally, but the interiors of the chassis and RF box were very clean and looked almost new. It had not been modified in any way. The SP-210, on the other hand, was in better condition on its topside surfaces, but was filthy underneath the chassis and had been extensively modified by a previous owner in a somewhat unprofessional manner. Both sets worked properly in all respects when first turned on. Excessive 120 Hz hum was traced to a defective filter capacitor in the external power supply, which otherwise also operated properly. New tubes were installed in the BC-779A unit and a complete alignment of the receiver was performed in accordance with instructions listed in the receiver technical manual [Ref. 3]. Sensitivity of the receiver was measured as a 10 dB signal+noise to noise ratio over the high frequency bands and the results are listed in Table I.

TABLE I. BC-779A SENSITIVITY MEASUREMENTS

Band (MHz)	Frequency (MHz)	S+N/N (microvolts)
2.5 - 5.0	2.5	16.0
2.5 - 5.0	5.0	5.9
5.0 - 10.0	5.0	8.5
5.0 - 10.0	10.0	4.0
10.0 - 20.0	10.0	14.0
10.0 - 20.0	20.0	4.3

Sensitivity of the SP-210 after similar alignment was somewhat better, but the extent of the previous modifications suggested that the BC-779A measurements would be a more valid standard of comparison. A design objective of the solid-state SUPER-PRO, therefore, was to match or exceed the specifications listed in Table I. Another characteristic of the BC-779A noted was a rapid and long-lasting drift in frequency toward the lower frequencies. This drift was finally reduced by cleaning the bandswitch contacts, but it was never completely checked and would have made the receiver unusable for sustained CW or SSB reception. Both receivers were also extremely sensitive to vibration and shock. The slightest bump or knock would cause them to shift noticeably in frequency.

As the SP-210 was already somewhat modified, that receiver was selected for solid-state conversion. The first step was the removal of the four tube audio circuitry. It was initially replaced with an interim audio amplifier using an integrated circuit capable of 2 watts audio output. This step permitted immediate removal of the +385 volt plate supply line and its attendant shock hazard. The interim

audio amplifier was mounted on an etched circuit board designed by the writer. The next stages intended for replacement were the BFO and detector-noise limiter. The product detector, however, was considerably delayed owing to a defective integrated circuit. While this trouble was being resolved, replacement of the front end stages began with the RF amplifiers. Both circuits worked properly on the breadboard and later on the single etched circuit board they were mounted on. The mixer and local oscillator stages were then developed the same way and were mounted on a second etched circuit board. Both front end circuit boards were mounted atop the chassis alongside the RF box. At this point, the product detector/AM detector and BFO were installed on completed circuit boards replacing the detector, noise limiter, and BFO tubes. After some amount of "de-bugging," the new stages operated in a satisfactory manner. At this point, all remaining vacuum tubes were removed and all unnecessary wiring was taken out. While the IF strip was being built on the breadboard, a new chassis made of aluminum was completed and the existing circuit boards were transferred to the new chassis, one by one. First installed was the completed audio module. The product detector board and BFO board were next, followed by the AGC and S-meter amplifier board and the marker generator board. The IF strip circuit board was completed and installed at this point. Further installations were delayed by trouble-shooting efforts in the IF amplifiers and fitting of a new front panel made to accommodate the larger number of controls. When the RF box was fitted into the new chassis, the front-end circuit boards were re-installed and the noise blanker circuit board was added shortly thereafter. The FM detector board and power supply board were added later as check-out efforts proceeded. Many of these stages were first constructed on a breadboard for check-out and experimentation. Others were proven circuits taken directly from manufacturer's application books, periodic literature,

or standard reference handbooks of recent publication.

Etched circuit boards were designed by the writer and constructed in the NPS Etching Laboratory using conventional photo-etch procedures. The boards were then tin-plated to enhance solder flow and to preserve the copper foil. The boards were subsequently drilled with small bits (sizes #60 and smaller) and assembled.

The power transformer was wound by the author with the aid of an employee and facilities of Magnetic Circuit Elements, Inc. of Monterey, California. The construction process was as follows: The specifications were established, the windings were placed on the plastic bobbin (the primary winding innermost,) insulating tape was wrapped over the completed windings, iron core material was stacked into the bobbin, a mixture of liquid plastic and thinner was brushed over the core material, and finally the frame structure was fitted over the assembled transformer. Filter chokes were also provided by Magnetic Circuit Elements, Inc.

As previously mentioned, a new chassis and front panel were built for the project. In addition, a rear panel and enclosure were similarly constructed. These items were laid out in scaled engineering drawings which were provided to the NPS Machine Facility where the items were constructed. The engraving of labels on the front panel was also performed at the Machine Facility. Rub-on labels were used on the rear panel after painting. Aluminum was used to effect a substantial weight savings over the original SP-210, which utilized a steel chassis and enclosure. (The BC-779A also had a steel front panel, making it prodigiously heavy.) An aluminum enclosure for the crystal filter replaced the original steel one, but no attempt was made to replace the steel construction of the massive RF box wherein lies most of the weight of the newly modified receiver.

Handles were mounted to the front panel to facilitate lifting and carrying the unit.

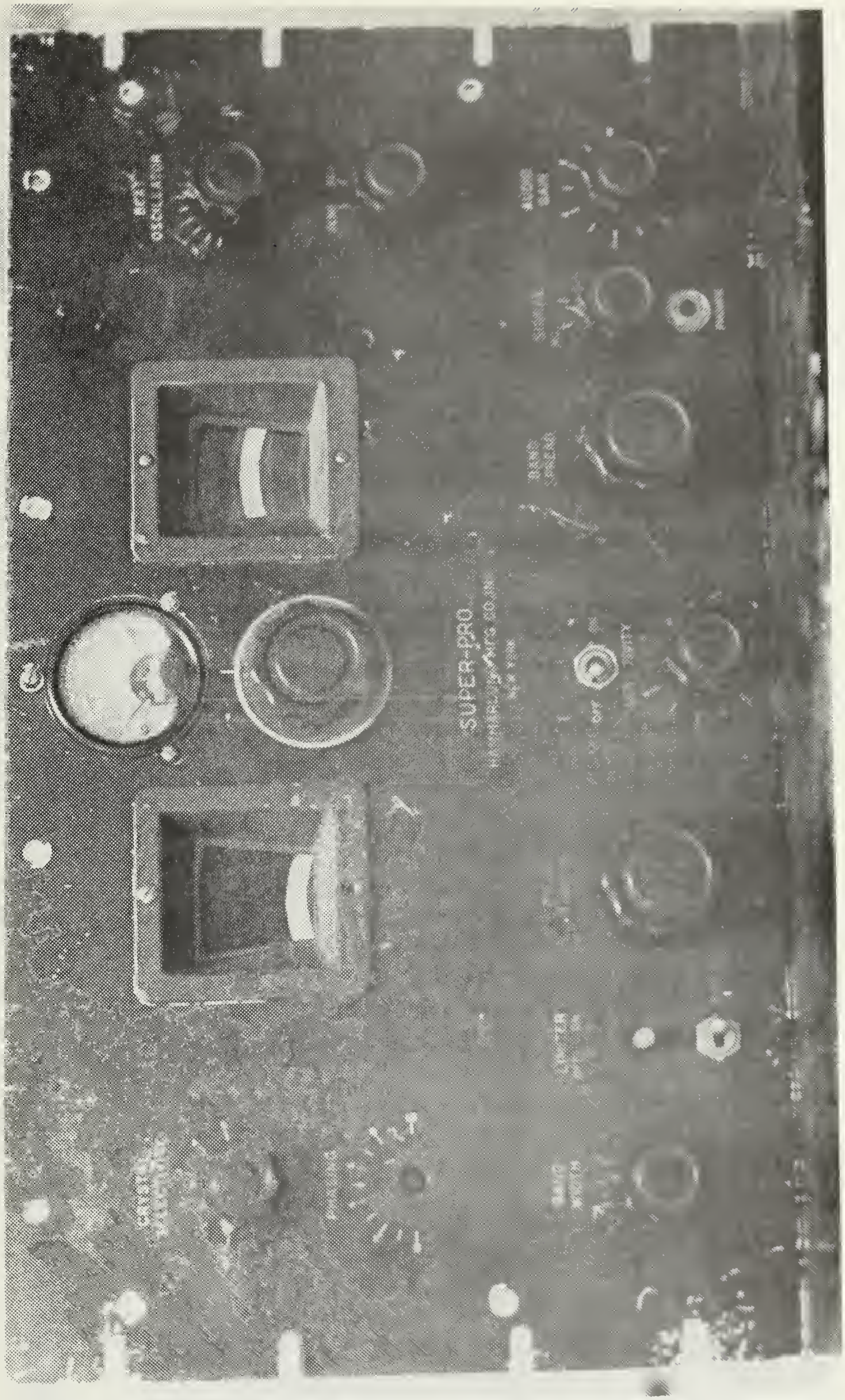

FIG. 4: FRONT PANEL: BEFORE MODIFICATION

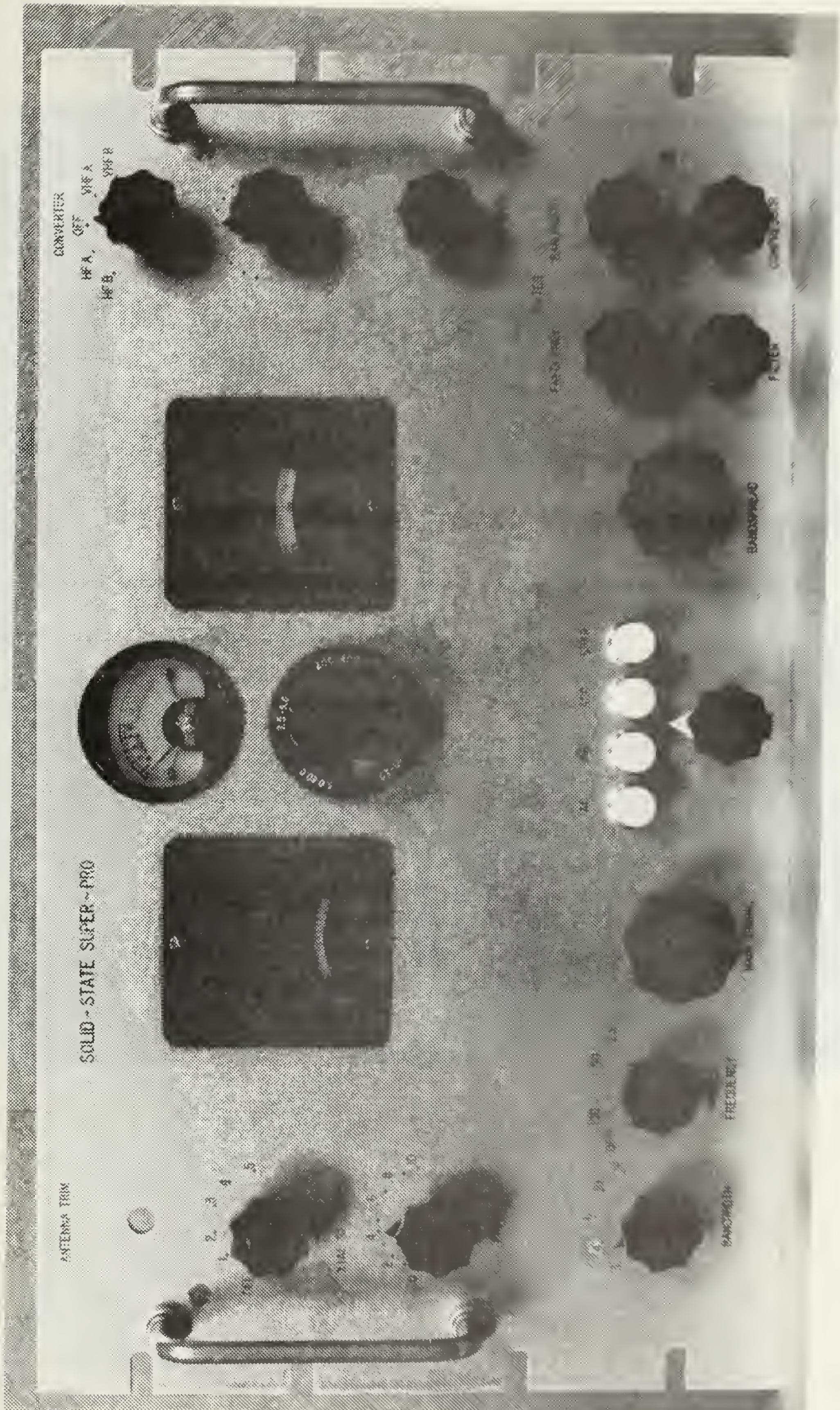

FIG. 5: FRONT PANEL: AFTER MODIFICATION

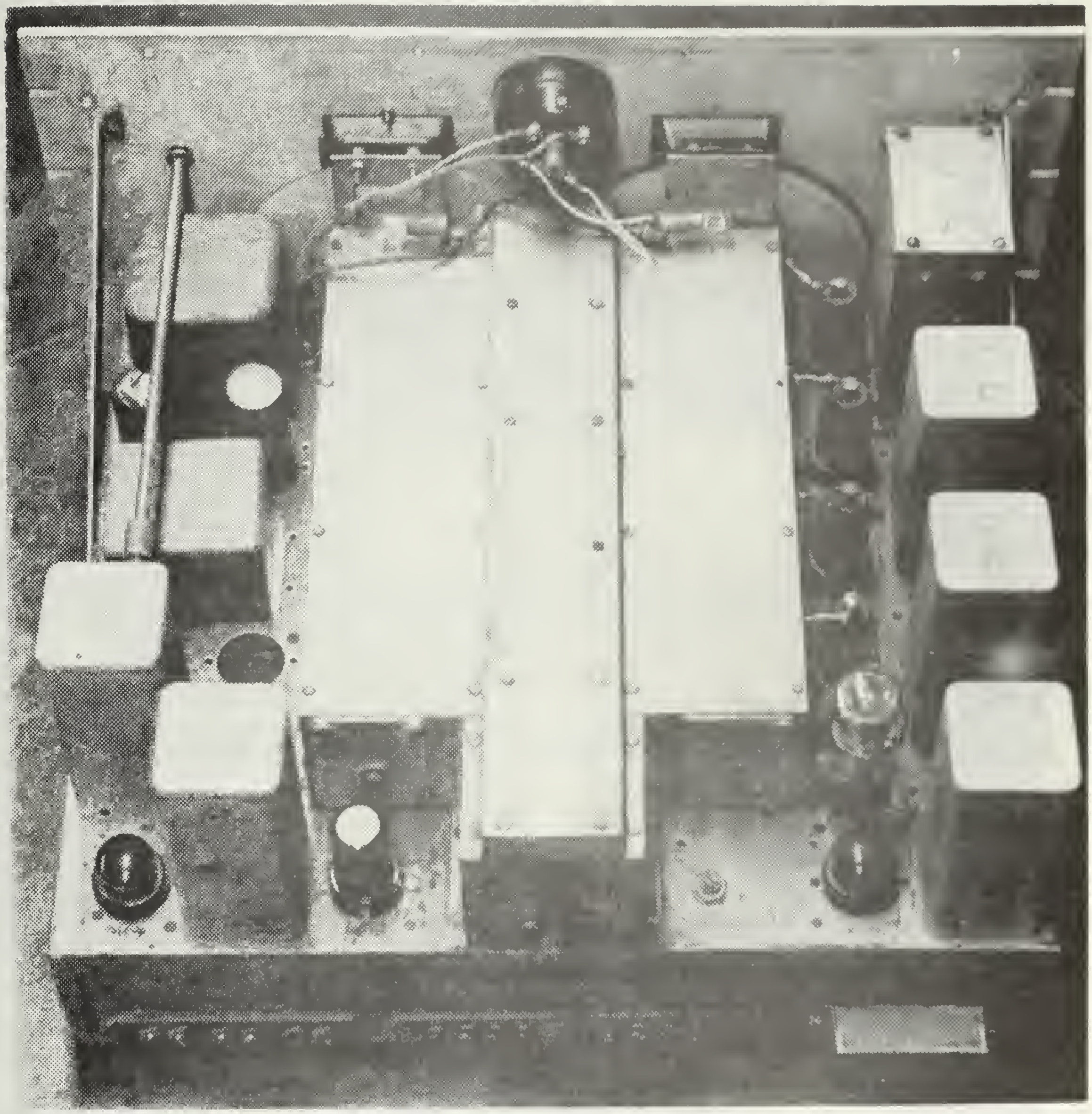

FIG. 6: TOP VIEW: BEFORE MODIFICATION

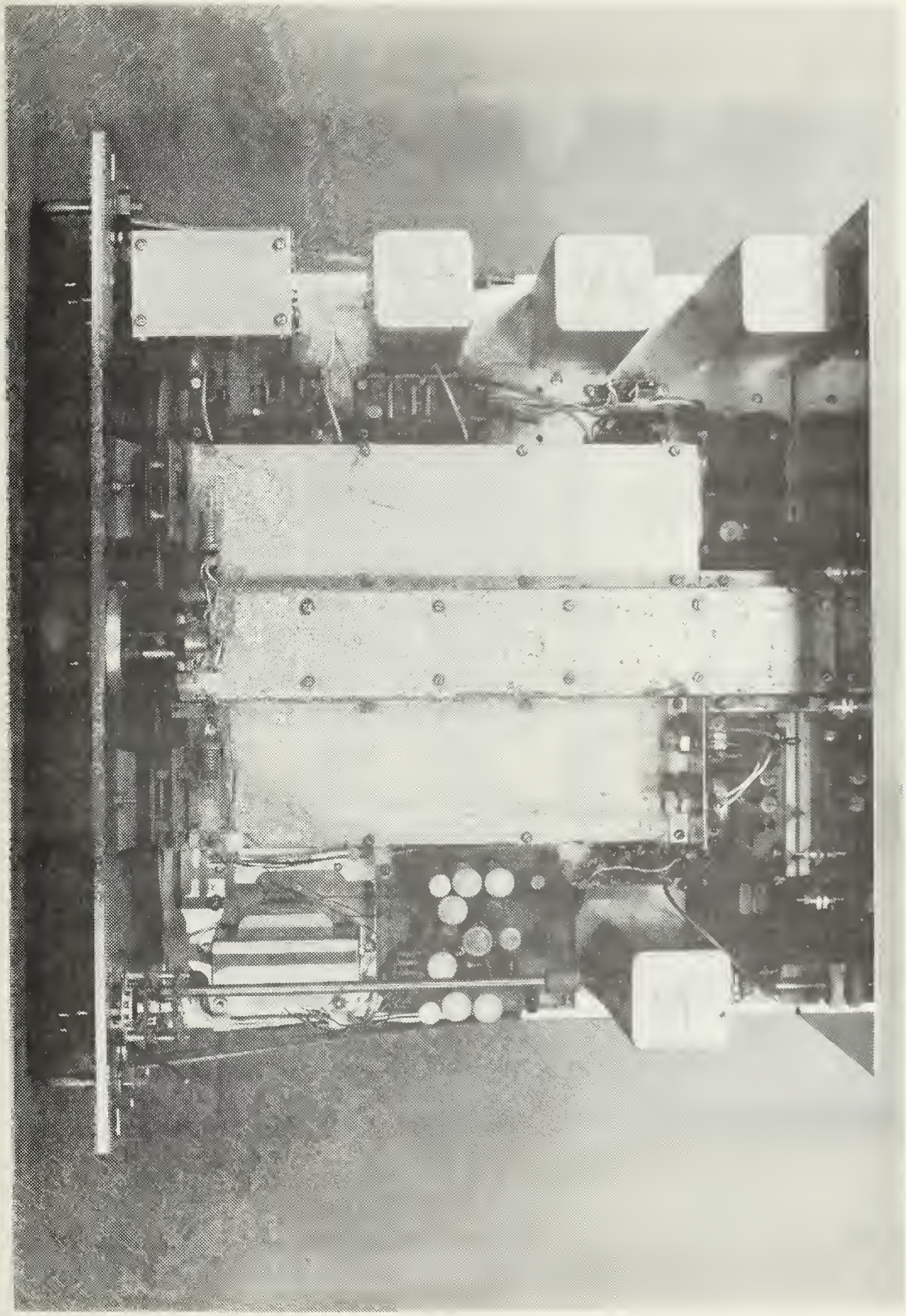

FIG. 7: TOP VIEW: AFTER MODIFICATION

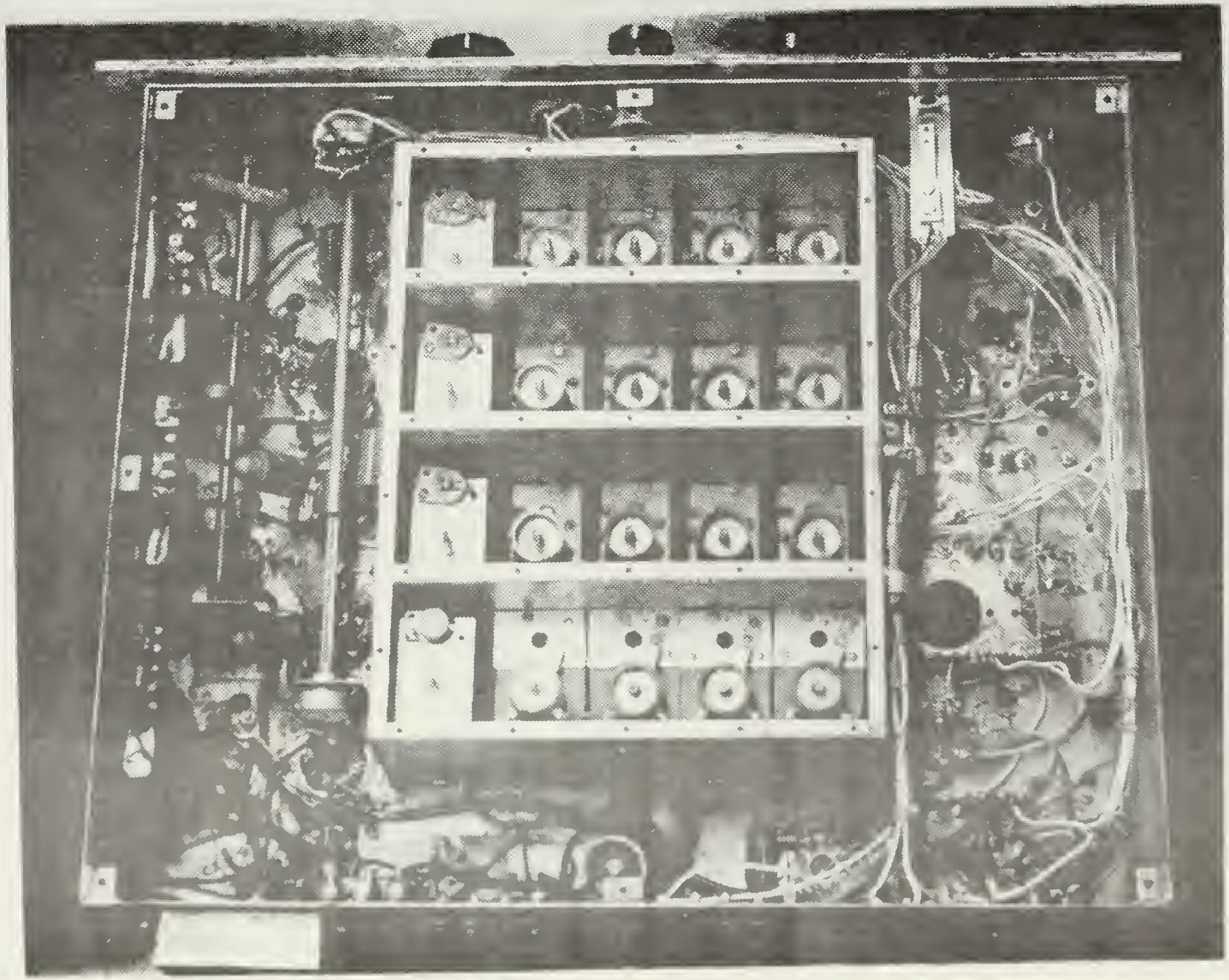

FIG. 8: BOTTOM VIEW: BEFORE MODIFICATION

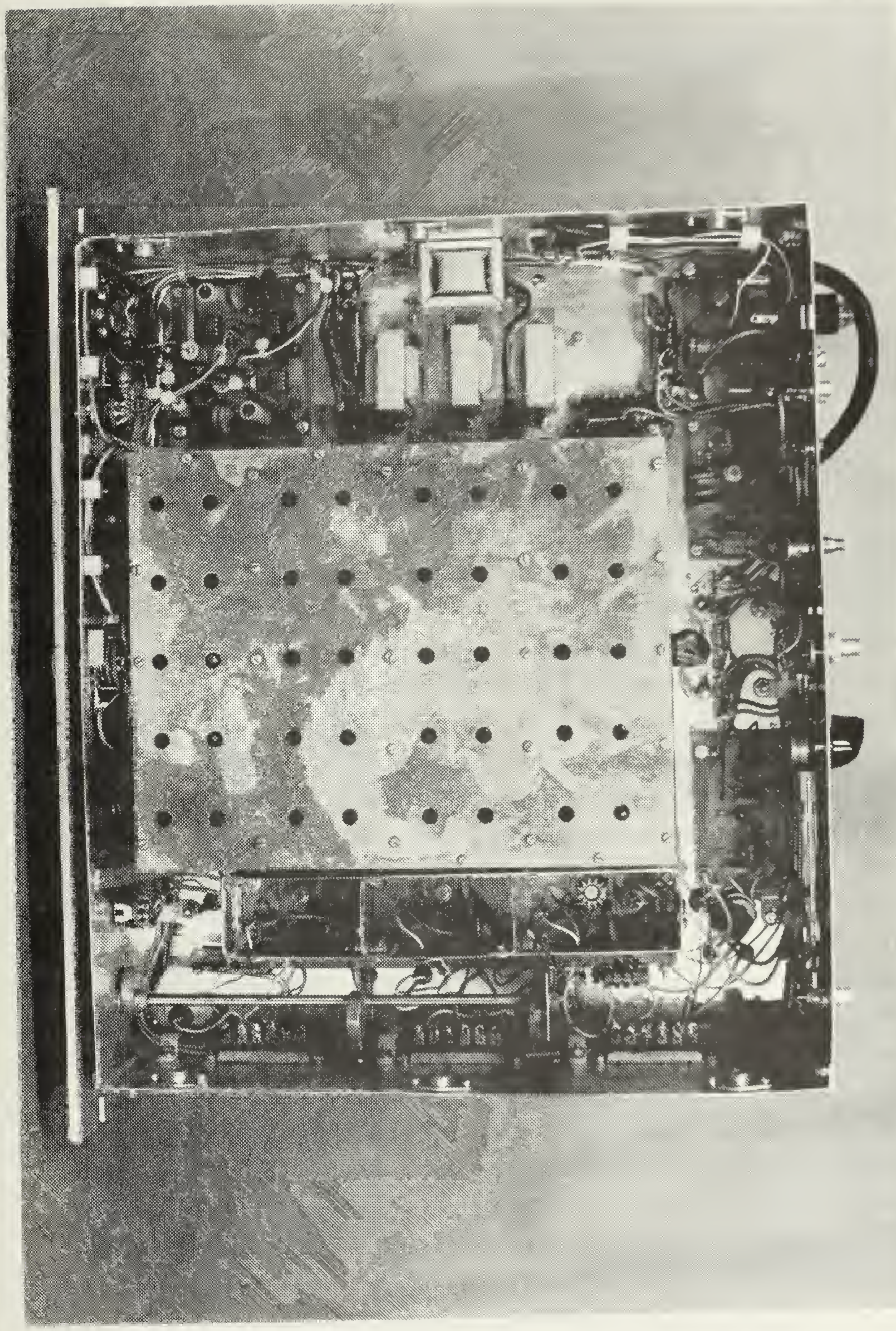

FIG. 9: BOTTOM VIEW: AFTER MODIFICATION

II. RECEIVER STAGE CONVERSION

A. CRYSTAL MARKER GENERATOR

Although there was no corresponding stage in the original design, a crystal marker generator was clearly needed. The rough calibration of the main tuning dial and the zero-to-100 logging scale on the bandspread dial do not allow precision frequency selection or measurement. The circuit chosen for this purpose, shown in Fig. 10, uses two digital integrated circuits. The fundamental calibrator frequency is set by crystal Y2 at 100 kHz. A multivibrator consisting of two NOR gates in U1, a Motorola MC724P quad gate, oscillates at the fundamental frequency. A third gate, with its inputs tied together, serves as a buffer and squarer to assure rich harmonic content of the 100 kHz. output. The fourth gate of U1 is not used. This output is fed to both S2A and U2, the latter a Motorola MC790P dual J-K flip-flop. The two flip-flops are connected as divide-by-two circuits and placed in cascade. The output of each flip-flop is also delivered to S2A where the operator may select markers at 100 kHz., 50 kHz., or 25 kHz. intervals or may switch off the generator. Switch S1B provides operating voltages to the generator unit and S1 as a whole is the CALIBRATOR control on the front panel, as shown in Fig. 5. The generator square wave output is coupled through C103, a 24 picofarad capacitor to the antenna input. C103 attenuates the stronger lower frequency harmonics to provide a more uniform input to the receiver over its wide range of frequency coverage. Alignment of the calibrator was accomplished by zero beat adjustment of the

trimmer capacitor, C102, with the receiver tuned to WWV. The marker generator, described by Blakeslee [Ref. 2], worked properly when first tested and required no further attention.

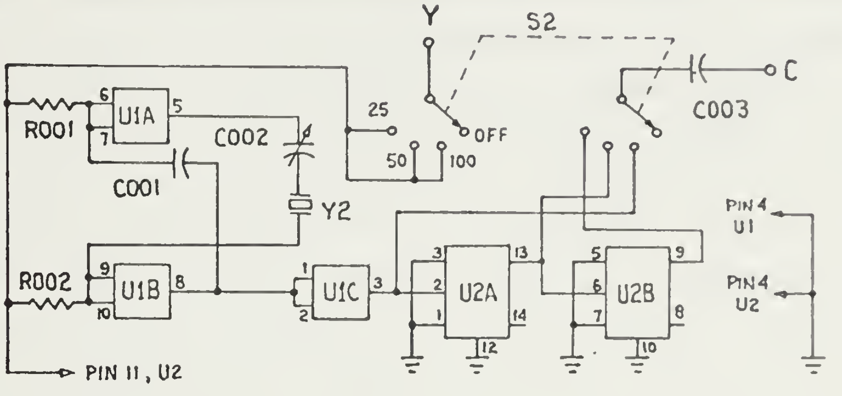

FIG. 10: MARKER GENERATOR

B. RADIO FREQUENCY AMPLIFIER STAGES

1. Original RF Amplifier Stages

Figure 11 shows the schematic diagram of the first RF stage. For simplicity, the bandswitching scheme has been omitted and only tuned circuit components of the 2.5 to 5.0 MHz. band were represented. Coil L3 in the tuning box was connected directly to the antenna terminals and was intended for use with a balanced type of antenna such as a dipole. Input impedance was approximately 115 ohms. An electrostatic shield was mounted between inductors L3 and L8 (as well as on coil pairs of the other bands) to minimize and fix capacitive coupling. Coil L8 was shunted by variable capacitors C61, C1A, and C2A, all located within

the tuning box. During RF alignment, L8 was tuned to the low frequency end of the band by means of a copper disc on a threaded shaft which was screwdriver adjustable from below the chassis. C61 was a trimmer which allowed proper alignment of the high frequency end of the band and was also screwdriver adjustable from below the chassis. Capacitors C1A and C2A were the first sections of the main tuning and bandspread controls, respectively. These controls were mounted on the front panel of the receiver. The 6K7 pentode amplifier tube was operated in a conventional grounded cathode circuit and represented a compromise between noise figure and gain. The pentode offered relatively high amplification at the expense of a higher noise figure than triode RF amplifiers of similar design. Output of the tube returned to the tuning box and went to the primary of L13.

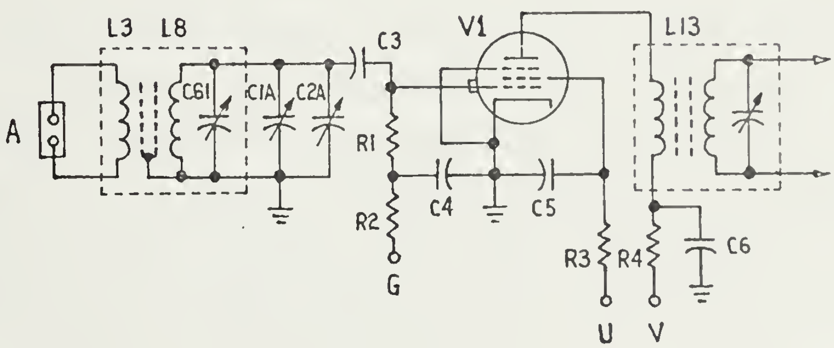

FIG. 11: ORIGINAL FIRST RF AMPLIFIER

The second RF amplifier, shown in Fig. 12, was almost identical to the preceeding stage. Again, the switching scheme was omitted in the diagram and components for the 2.5 to 5.0 MHz. band only are shown. Inductor L13 and trimmer capacitor C66 were adjusted for proper alignment from below the chassis. Capacitors C1B and C2B provided main tuning and bandspread for the circuit. Output of the 6K7

amplifier tube was to inductor L18 in the tuning box.

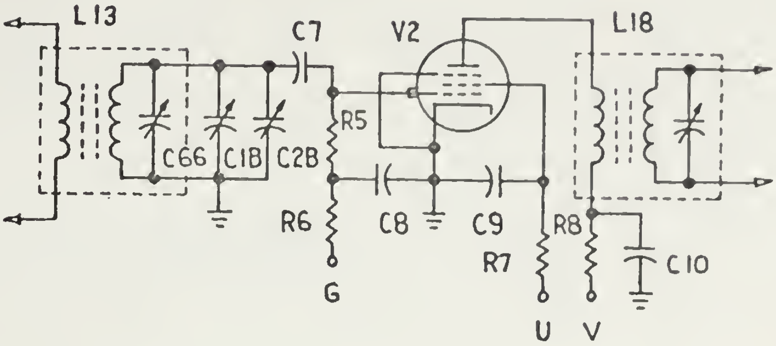

FIG. 12: ORIGINAL SECOND RF AMPLIFIER

2. Modified RF Amplifier Stages

The circuit shown in Fig. 13 is a modification of a circuit offered by Blakeslee [Ref. 2]. Two junction field-effect transistors are operated in a cascode circuit. This type of circuit normally requires neutralization at frequencies above 30 MHz., but since the SUPER-PRO does not tune frequencies higher than 20 MHz., neutralizing circuitry was not included. Experiment showed that none was required. AGC potential can be applied to the gate of transistor Q2, although experiment showed that sufficient AGC action was accomplished by the second RF amplifier and the IF strip. Therefore, the AGC capability of this circuit was not used. No difference in performance was observed with resistor R103 grounded or simply left unconnected and the latter arrangement was used. The transistors employed in the circuit are Motorola HEP802 field-effect transistors, which are commercial replacements for the type MPF102, also manufactured by Motorola and used elsewhere in this conversion.

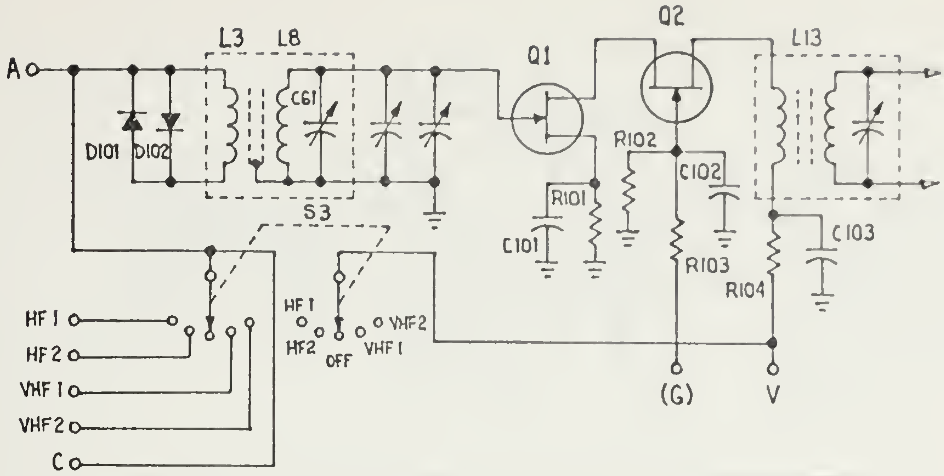

FIG. 13: MODIFIED FIRST RF AMPLIFIER

The second RF amplifier of the modified SUPER-PRO is shown schematically in Fig. 14. The heart of the amplifier is a Plessey SL610C RF/IF amplifier integrated circuit. Due to the low input impedance of the integrated circuit, a buffer stage using a 2N3705 junction transistor (Q3) in an emitter follower configuration was used to match the high impedance tuned circuit to the integrated circuit input. This measure was found to be necessary by experiment and values shown for the buffer circuit were those which allowed the largest dynamic range of input signal with minimum distortion for the specific transistor selected. The remainder of the circuit was taken from the Plessey integrated circuit application manual [Ref. 4]. AGC potential is provided to pin 7 of the integrated circuit. The Plessey device was specifically selected for this stage because of its AGC characteristics when teamed with a second Plessey integrated circuit, the SL621C AGC generator. The SL621C appears in this conversion as U13 and is described in Paragraph II.I.2 below.

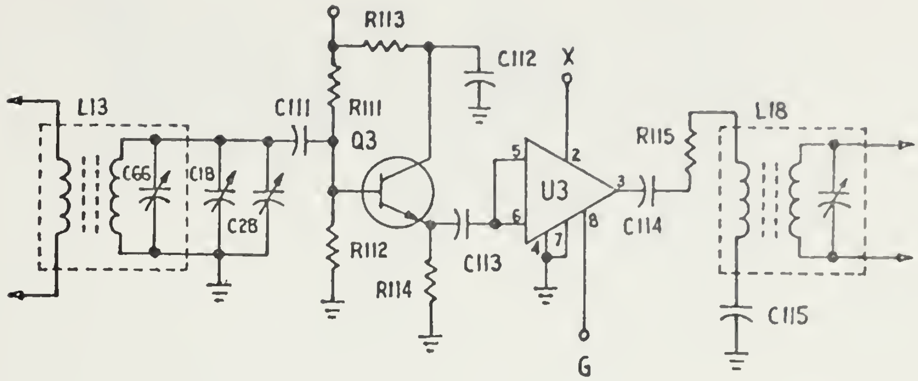

FIG. 14: MODIFIED SECOND RF AMPLIFIER

C. MIXER STAGE

1. Original Mixer Stage

The original SUPER-PRO mixer was a standard circuit using a 6L7 pentagrid mixer tube as shown in Fig. 15 and only the 2.5 to 5.0 MHz tuned circuit components are shown. Inductor L18 and trimmer capacitor C71 tuned in a similar manner to the RF amplifier tuned circuits. C1C and C2C were main tuning and bandspread variable capacitors respectively. Mixing action combined the RF and local oscillator signals and produced the standard mixer products. Output was to T1, the crystal filter and associated circuitry, which accepted the difference frequency (465 kHz) and rejected the other mixer products.

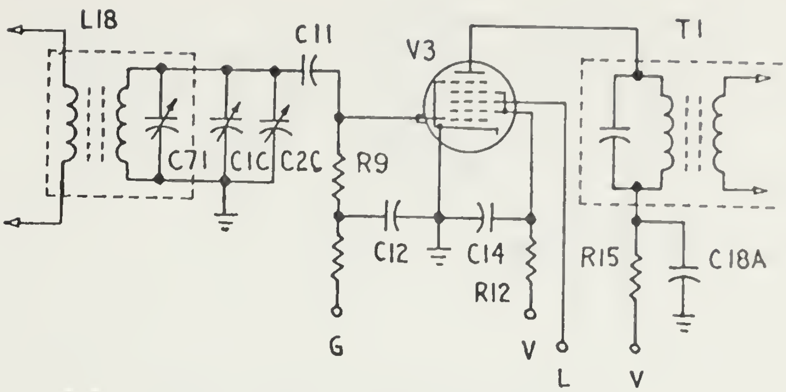

FIG. 15: ORIGINAL MIXER

2. <u>Modified Mixer Stage</u>

The new mixer, shown in Fig. 16, uses the same input and output tuned circuits, including the original C11. The bandswitching circuits have also been retained. Replacing the 6L7 mixer tube is a zener-protected dual-gate metal-oxide substrate field-effect transistor (MCSFET) manufactured by Motorola. Specifically advertised as a mixer, the MPF122 is installed with the RF input to gate 1 and the local oscillator input to gate 2. The circuit offers excellent immunity to cross-modulation and overload. The isolation effected between the local oscillator and the RF stages is considerably better than that of other types of discrete transistor mixers. In addition, the high input impedance of both transistor gates makes this circuit more easily suitable for replacement of the vacuum tube mixer than either a junction transistor mixer or an integrated circuit balanced mixer, such as the Motorola MC1496G or equivalents.

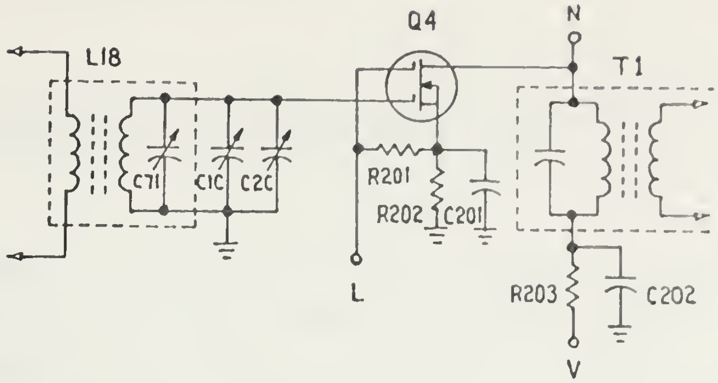

FIG. 16: MODIFIED MIXER

Problems in the mixer stage involved the transistor Q4 and the tuned circuits. Supplied in a non-standard and unidentifiable package, the MPF122 MOSFET had to be tested in operation to determine lead identity. An incorrect evaluation of this test enabled the author to prove that improperly installed MOSFET mixers are very lossy devices. The tuned circuits were unable to resonate properly owing to increased shunt capacitance added by Q4. The removal of a few turns of wire from inductor L23 improved the tuning characteristics in the 2.5 to 5.0 MHz. frequency band, but did not provide wholly satisfactory results. Point N in Fig. 16 is the sampling port for the noise blanker stage described in Paragraph II.G.3 below. Point L is the local oscillator input and Point V is the +12 VDC line.

D. LOCAL OSCILLATOR STAGE

1. Original Local Oscillator Stage

Shown in Fig. 17 is the original local oscillator of the SUPER-PRO receiver. The oscillator tube, a 6J7 pentode, was operated as a triode with its screen and suppressor grids tied externally to the plate. The circuit itself was a Hartley oscillator. As in other front end circuits previously described, the bandswitching circuits were omitted in Fig. 17 and only those tuned circuits for the 2.5 to 5.0 MHz band were shown. Frequency alignment of the local oscillator stage was accomplished by tuning the trimmer capacitor C76 to the high frequency end of the band and adjusting inductor L23 to the low frequency end. Capacitor C81 was a fixed padder and the ganged variable capacitors C1D and C2D were main tuning and bandspread controls, respectively. The oscillator was tuned to a frequency 465 kHz higher than the desired reception frequency so that the mixer output was equal to the intermediate frequency of 465 kHz. Output was coupled through C13, a 95 pf capacitor to the injection grid of the mixer tube. The amplitude of the local oscillator signal was about 1 volt RMS at 5 MHz.

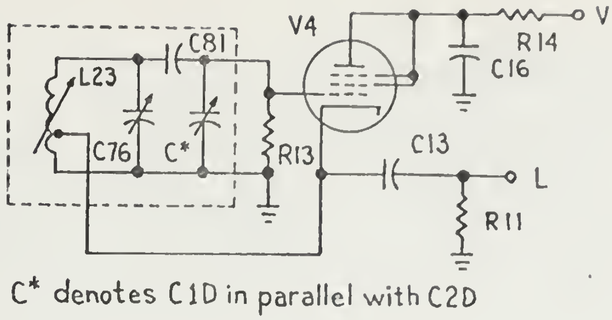

FIG. 17: ORIGINAL LOCAL OSCILLATOR

2. Modified Local Oscillator Stage

The new local oscillator stage, illustrated in Fig 18, uses a Motorola MPF102 field-effect transistor (FET) oscillator in a Hartley configuration, followed by an RCA 40673 dual gate MOSFET buffer stage. As in other front end stages, the bandswitching and tuning systems have been retained without modification. The diode D211 in the gate circuit of Q5 clamps the positive-going half of each cycle to prevent Q5 from reaching its high peak transconductance, the time period when oscillator output is rich in harmonic energy. Blakeslee [Ref. 2] recommends this addition to oscillators using either junction or MOS field-effect transistors. Output of Q5 is taken from the source and is fed to the gates of Q6 which are tied together. Q6 is operated as a source follower amplifier and therefore offers a voltage gain slightly less than unity. The advantage of using Q6 in this configuration is the constant load it places on Q5, a factor which enhances frequency stability of the local oscillator signal and thus reduces the drift characteristic of the unmodified receiver. The amplitude of the local oscillator output is approximately 0.2 volts RMS.

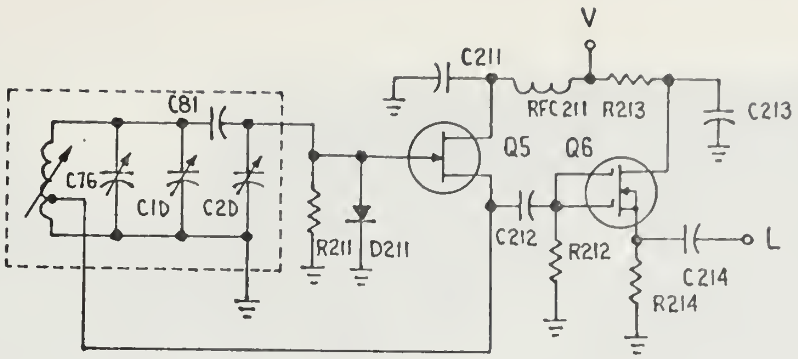

FIG. 18: MODIFIED LOCAL OSCILLATOR

The local oscillator operated satisfactorily from the first breadboard trial to final assembly and installation in the modified receiver chassis. No significant troubles were encountered with the new local oscillator circuitry. Difficulty was, however, experienced in tuning the original tuned circuits to resonance and this problem was partially solved in the same manner as that described for the mixer stage.

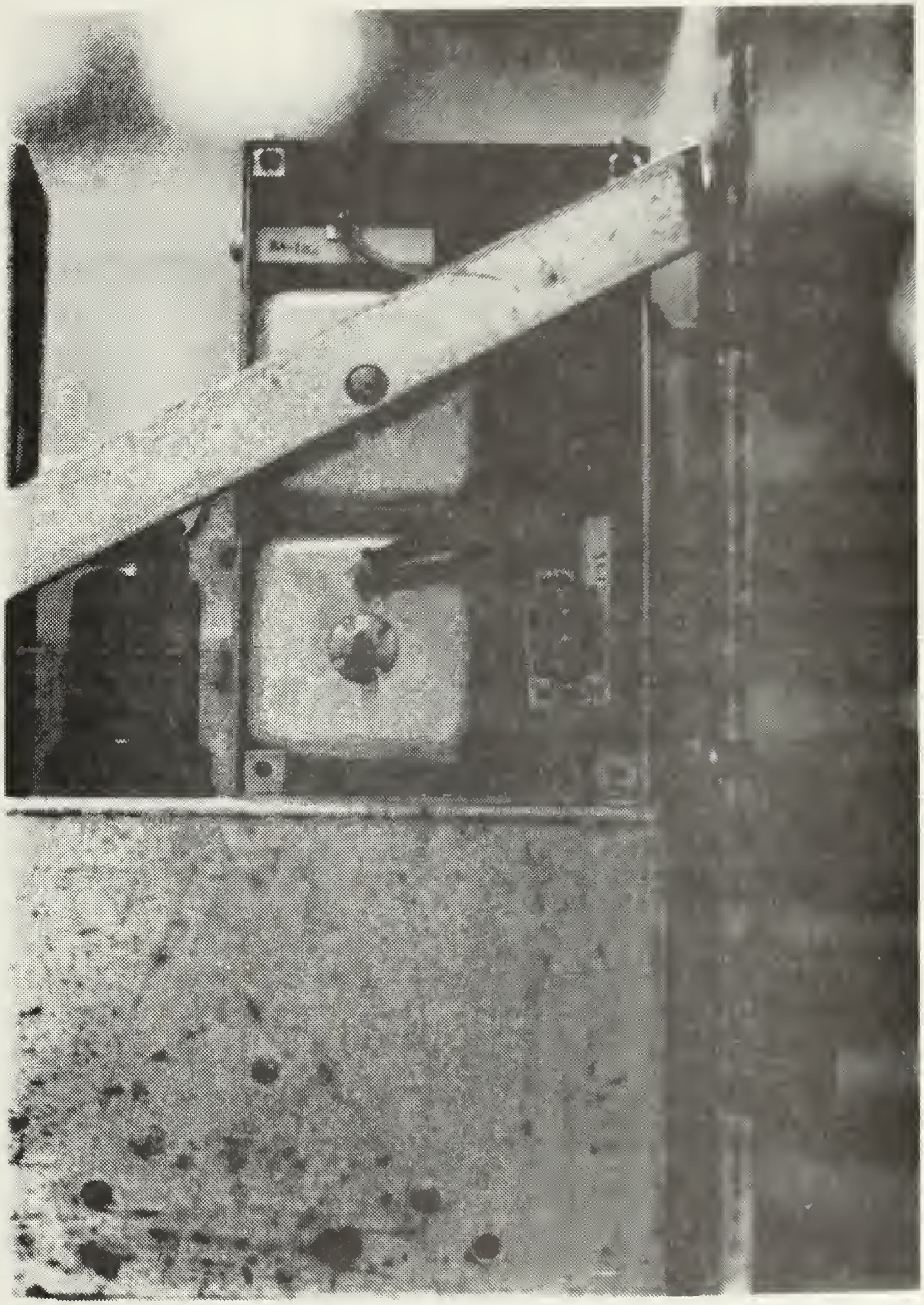

FIG. 19: CRYSTAL FILTER IN ORIGINAL SUPER-PRO

E. CRYSTAL FILTER

The crystal filter in the SUPER-PRO was one of the first such devices and was intended to enhance receiver IF selectivity. It was a simple, but ingenious circuit making use of only one piezoelectric crystal. For simplicity, the crystal filter unit was designated "T1." Figure 19 is a photograph of the original T1 enclosure with the cover removed and the complete schematic diagram of the filter is shown in Fig. 20. This stage survived the modernization with no significant modification.

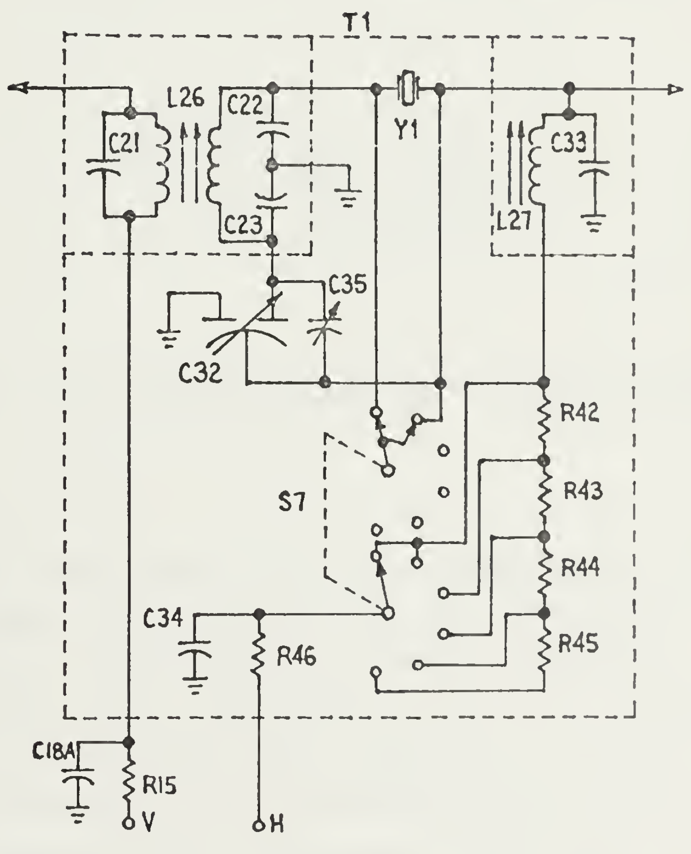

FIG. 20: ORIGINAL CRYSTAL FILTER T1

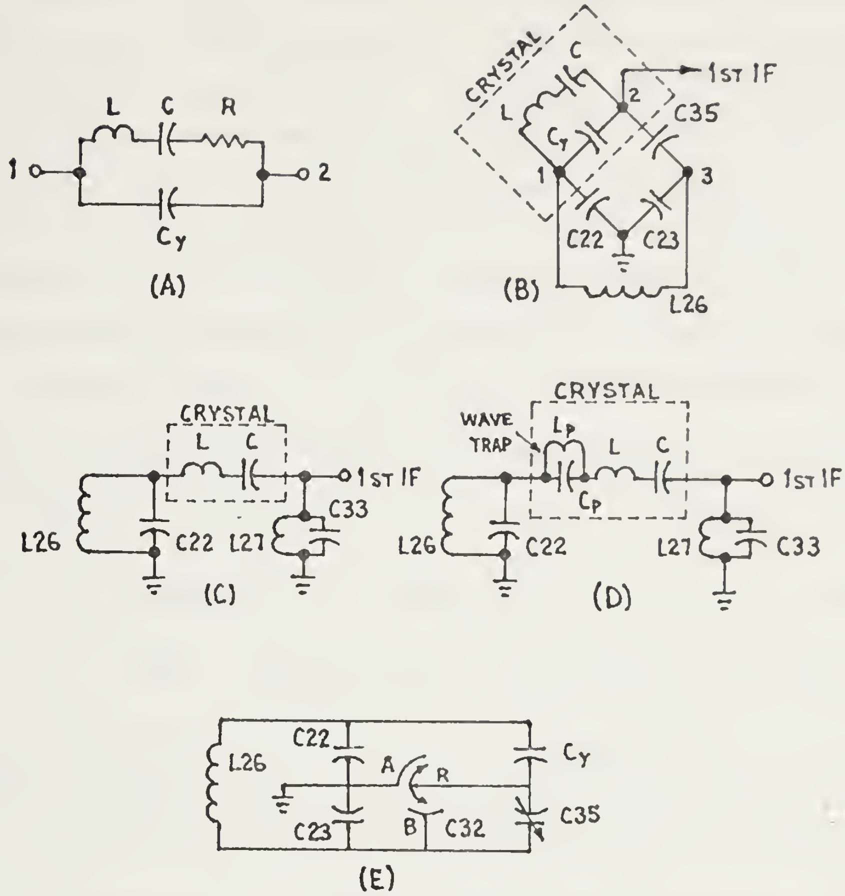

FIG. 21: CRYSTAL FILTER EQUIVALENT CIRCUIT

Figure 21A showns the electrical equivalent circuit of the crystal Y1. This circuit was resonant at either of two frequencies. The series resonant frequency was that frequency where the reactance of L was equal to the reactance of C. The parallel resonant frequency was the frequency at which the inductive reactance of the L-C series combination equaled the capacitive reactance of C_y. At series resonance, the crystal represented a low impedance.

At parallel resonance, it was a high impedance element.

Figure 21B shows a simplified circuit which represents basic filter action. L26 was the secondary of the IF input transformer. Across this inductor was a capacity bridge composed of C22 and C23, two fixed 100 pf mica capacitors, C35, a 5 pf trimmer capacitor, and crystal Y1, which is again represented by its equivalent circuit. The junction of C22 and C23 was grounded and the output was taken from the junction of C35 and Y1. When the capacitance of C35 equaled that of C_y, the voltage coupled from point C to point B through C35 was equal and opposite to the voltage coupled from point A to point B through C_y. This effectively cancelled the effect of C_y and Y1 was series resonant at the intermediate frequency, 465 kHz. Under these conditions, the simplified circuit of T1 reduced to that shown in Fig. 21C. If the capacitance of C35 was less than C_y, C_y was no longer neutralized and Y1 operated in a parallel resonant manner with the resonant frequency slightly greater than the intermediate frequency. The simplified circuit was then as shown in Fig. 21D. L_p and C_p attenuated signals at the parallel resonant frequency, but ignored those at the intermediate frequency. Similar action occurred when the capacitance of C35 was greater than C_y, except that the parallel resonant frequency was slightly less than the intermediate frequency. Thus signals at frequencies below the intermediate frequency were attenuated while those at the intermediate frequency were not. C35 was carefully set to the exact capacitance of C_y during receiver

alignment to prevent detuning of T1 from the intermediate frequency. In order to shift the filter rejection frequency to either side of the intermediate frequency, the differential phasing capacitor C32 was placed in the capacitor bridge as shown in Fig. 21E. An air variable capacitor, C32 has two stator sections so arranged that as the rotor turned out of one stator section, it turned into the other by an equal amount. In this way, the capacitance of one section increased as the capacitance of the other section decreased. Only one of the stator sections was in parallel with C35, however, and by placing more or less capacitance in parallel with C35, the system tuned the parallel resonant trap of Fig. 21D to frequencies as much as a few kilohertz either side of the intermediate frequency. This system also maintained a constant capacitance across L26 and thus did not detune the tuned circuit of the IF input transformer. When C32 was set at its mid-point, the equivalent circuit reverted to Fig. 21C. Capacitor C32 was the PHASING control on the front panel as shown in Fig. 4.

Figure 20 shows the complete circuit of T1. In the "OFF" position, the CRYSTAL SELECTIVITY switch S7 shorted Y1, reducing T1 to a conventional IF transformer. In position 1, S7 inserted the crystal to the circuit as described above. The parallel resonant grid circuit L27-C23 prevented loss of the signal in the grid bias path and tuned the grid of the first IF amplifier tube. Advancement of S7 to positions 2 to 5 placed resistors in series with the L27-C23 tank circuit. This progressively increased the Q of the filter output circuit and the selectivity of the entire circuit approached that of the crystal alone.

In the rebuilt receiver, the only electronic modification to the filter was the removal of resistor R46, which established grid bias for V5, the first IF amplifier. The original phenolic shaft of capacitor C32 had been broken

off years before and a new one made of nylon was fitted. As previously mentioned, an aluminum enclosure replaced the original steel box containing the components of T1. The alignment of T1 necessitated the use of an IF sweeper generator and a triggered oscilloscope. A listing of all test equipment used for this project is contained in Appendix E.

F. INTERMEDIATE FREQUENCY AMPLIFIER STAGES

1. Original IF Amplifier Stages

The original SUPER-PRO IF amplifier strip used three pentode amplifier tubes in cascade with transformer coupling. The first stage used a 6K7 with input from T1 fed to the grid cap on the tube. Bias and AVC voltage was supplied to the tube via T1 circuitry. As shown in Fig. 22, the amplifier was a conventional grounded cathode arrangement with output from the V5 plate to the primary of T2.

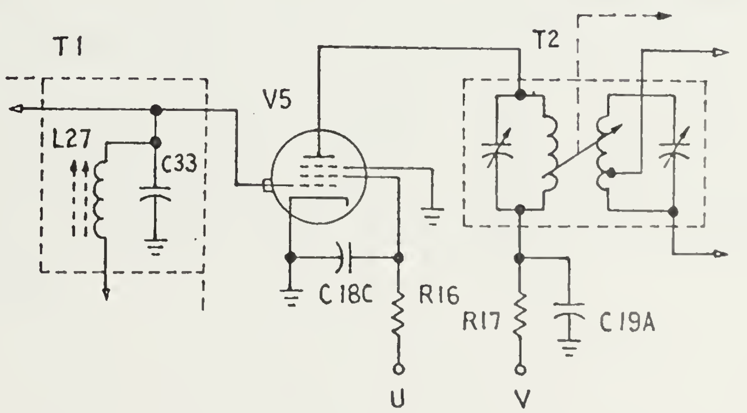

FIG. 22: ORIGINAL FIRST IF AMPLIFIER

The second IF amplifier, illustrated in Fig. 23, was very similar to the previous stage. AVC voltage was supplied via the bias resistor R18 and the primary cf T2. The amplifier tube V6 was a 6SK7, which had about the same operating characteristics as the 6K7 employed in the previous stage, but did not have a grid cap. The cutput went from the V6 plate to the primary of T3.

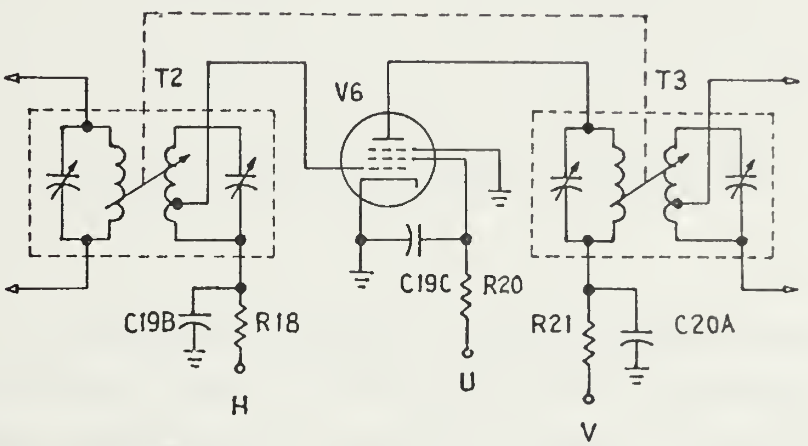

FIG. 23: ORIGINAL SECOND IF AMPLIFIER

Input to the third IF amplifier came from the preceeding stage via T3. As shown in Fig 24, a sample of this input was directed to the AVC tube V11. The third IF stage was operated with fixed bias, but was ctherwise similar to the preceeding stage. Output went to the primary winding of T4 where the BFO signal was injected through capacitor C41.

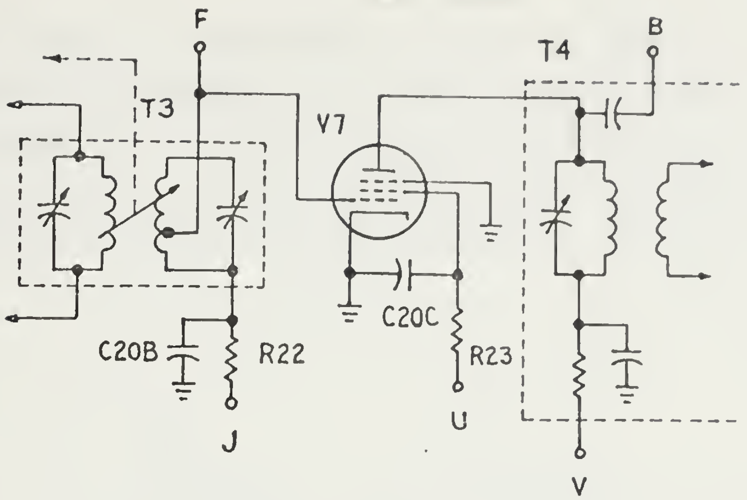

FIG. 24: ORIGINAL THIRD IF AMPLIFIER

An interesting feature of the receiver was the type of IF transformers used. Both primary and secondary of T2 and T3 were tuned by air vairable trimmer capacitors. The inductors were triple pie-wound coils whose coupling was varied by sliding decks on which the primary coils were mounted. A cam and lever assembly enabled the operator to vary the IF coupling of both T2 and T3 simultaneously from the front panel. This control was the BANDWIDTH control in Fig. 4.

2. Modified IF Amplifier Stages

The new IF amplifier strip is composed of three nearly identical circuits, each built around the Motorola MC1590G integrated circuit, whose internal circuit diagram is shown in Fig. 49 in Appendix D. This device is characterized by ample gain and excellent AGC response.

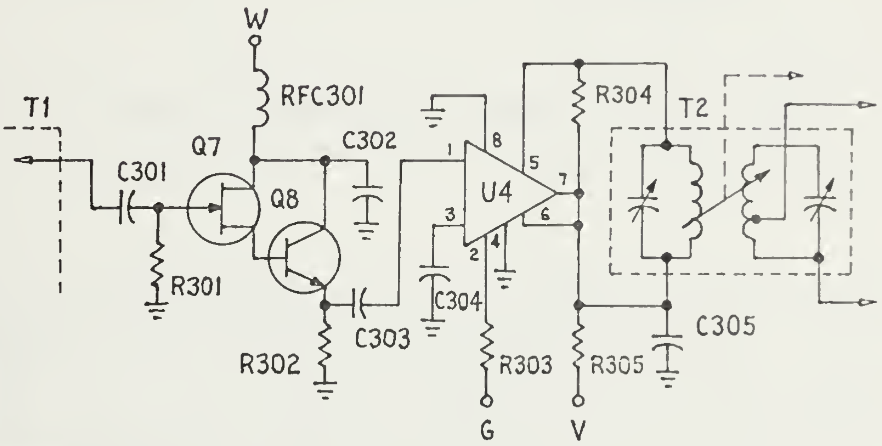

FIG. 25: MODIFIED FIRST IF AMPLIFIER

The first, second, and third IF amplifier circuits are shown in Figs. 25, 26, and 27, respectively. Resistors R304, R314, and R324, across the integrated circuit output leads, serve as 1000 ohm resistive loads and reduce gain to a conservative, but still adequate level. Without these resistors, the circuits oscillated due to the high load impedance presented by the IF transformer primary windings.

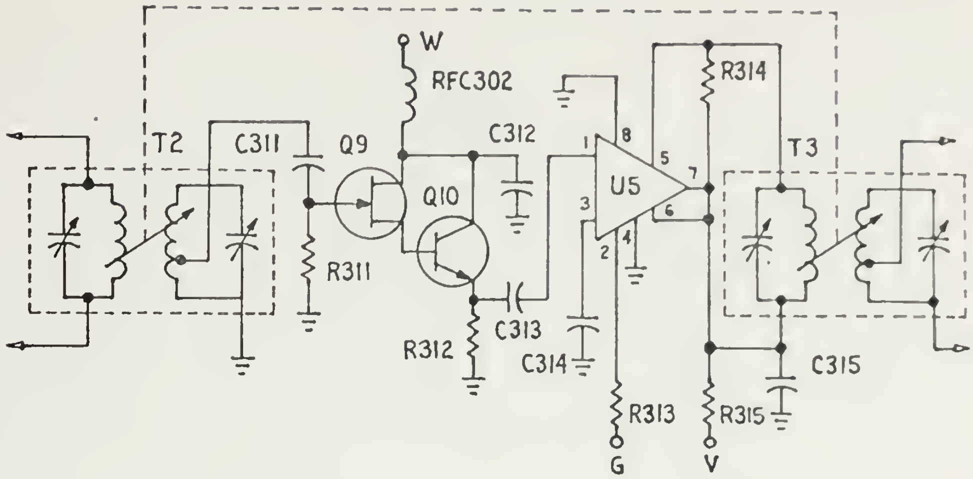

FIG. 26: MODIFIED SECOND IF AMPLIFIER

In each stage, a two transistor buffer provides a low impedance load on the secondary windings of the IF transformers. In a circuit offered by Eslick [Ref. 5], an FET source follower using an MPF102 is in cascade with a junction transistor emitter follower employing a 2N3706 transistor. This circuit provides excellent isolation between input and output and ample dynamic range for this application.

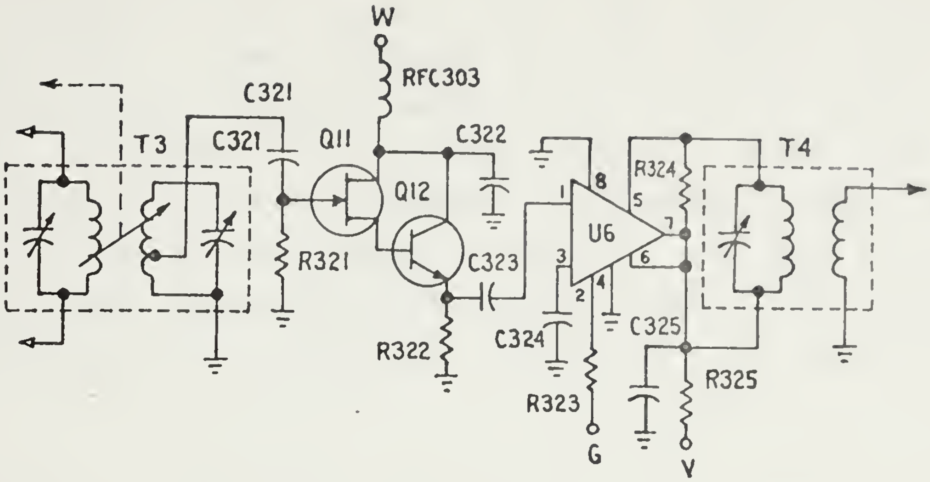

FIG. 27: MODIFIED THIRD IF AMPLIFIER

As expected, the IF strip offerred more problems than any other stage of the modified receiver. The integrated circuits were chosen as amplifiers over MOSFETs in order to gain experience with integrated circuits and because their positive voltage AGC characteristics promised good performance when combined with the Plessey SL621C integrated circuit AGC generator used in this conversion. Incompatibility with the IF transformers (a trouble not anticipated if MOSFET IF amplifiers had been used,) was foreseen and confirmed by experiment. Both before and after the circuits shown in Figs. 25, 26, and 27 were assembled on an etched circuit board, stray feedback, with resultant oscillation, plagued the IF amplifier strip. The temporary installation of simple shields between the amplifier stages prevented most of the oscillation occurrences. Therefore, an elaborate aluminum shield box was built around the IF strip circuit board. This step sacrificed the easy accessibility theretofore enjoyed with the unshielded circuit board, but further reduced oscillation tendencies.

The high gain of the amplifiers, even with resistors R304, R314, and R324 in place, made the IF strip very sensitive to signals injected at its input, but also quite unstable, especially with close coupling in T2 and T3. These IF transformers tuned satisfactorily only in the secondary circuits. Adjustment of primary trimmer capacitors had little effect on receiver response, suggesting significant impedance mismatch and circuit detuning. This was probably due to the load resistors R304, R314, and R324, without which, however, the circuits oscillated, as previously stated.

G. DETECTOR AND NOISE REDUCING STAGES

1. Original Detector Stage

The detector stage in the original SUPER-PRO was a simple slope detection affair using a 6H6 dual diode with its diode sections in parallel as shown in Fig 28. The signal was rectified in V8 and the audio was developed across R25 after the 50 pf capacitors C44, C45, and C26 filtered out RF components. Output of the detector went to R26, the audio gain control in the audio amplifier section.

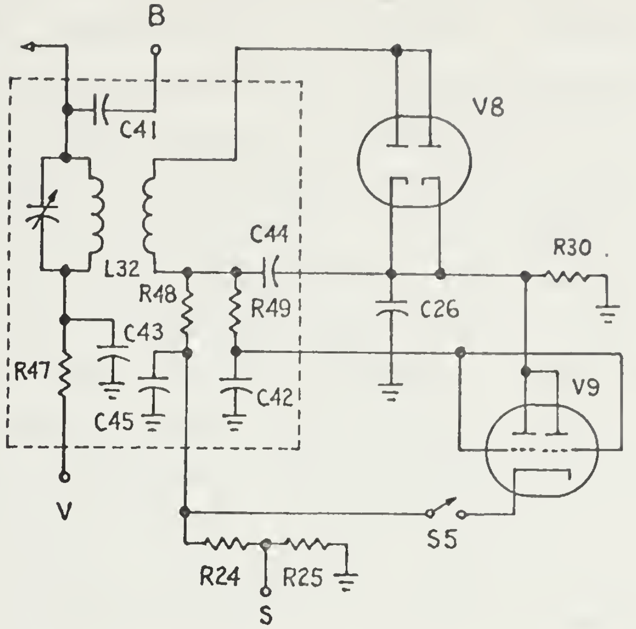

FIG. 28: ORIGINAL DETECTOR/LIMITER

2. Original Noise Limiter Stage

The noise limiter of the unmodified receiver followed the detector stage in the receiver block diagram, Fig. 2. Limiting action was effected by V9, a 6N7 dual triode connected so that its triode sections were in parallel. A four ohm resistor (R31) in the filament circuit of V9 reduced its filament voltage to about 3 volts. As shown in Fig. 28, the tube elements connected to various resistors in the detector output circuit. The plates were biased at the detector cathode voltage. V9 cathode voltage was drawn from a voltage divider in the V8 cathode circuit in such a way as to provide a constant voltage drop across V9 during constant amplitude signal conditions. The grids were negative with respect to the cathodes and the tube

normally operated at cut-off. A high amplitude noise spike increased the plate-to-cathode voltage drop. Capacitor C42 held the grids momentarily at their original voltages, causing the tube to conduct to a degree proportional to the amplitude of the noise spike. Conduction of V9 provided a low impedance path for the spike and it bypassed R24, across which the audio was developed. The time constant of R49-C42 in the V9 grid circuit controlled the time response of the limiter to signals of varying amplitude. For large noise pulses, it was too slow to restore cut-off bias to V9 and effectively shorted them to ground. It occasionally degraded highly amplitude modulated signals, but for slower variations, such as fading and casual tuning, the R-C time constant of R49-C42 allowed the charge on C42 to readjust itself rapidly enough to prevent V9 from coming out of its cut-off condition.

3. Modified Receiver Noise Blanker Stage

Unlike the original noise reduction circuit, which followed the detector stage, the new noise blanker is located in the IF strip. The original circuit was a limiter which clipped off noise pulses at the ambient audio level. The noise blanker circuit cuts off reception entirely during such a pulse and allows the operator to integrate mentally the portions of the recieved signal passed by the blanker before and after the pulse. In most cases, the operator does this automatically and does not perceive the gap in the received singal. If the signal is sampled in the mixer before noise pulses are broadened by tuned circuit action in the IF strip, gaps in the received signal are on the order of microseconds and are unnoticed.

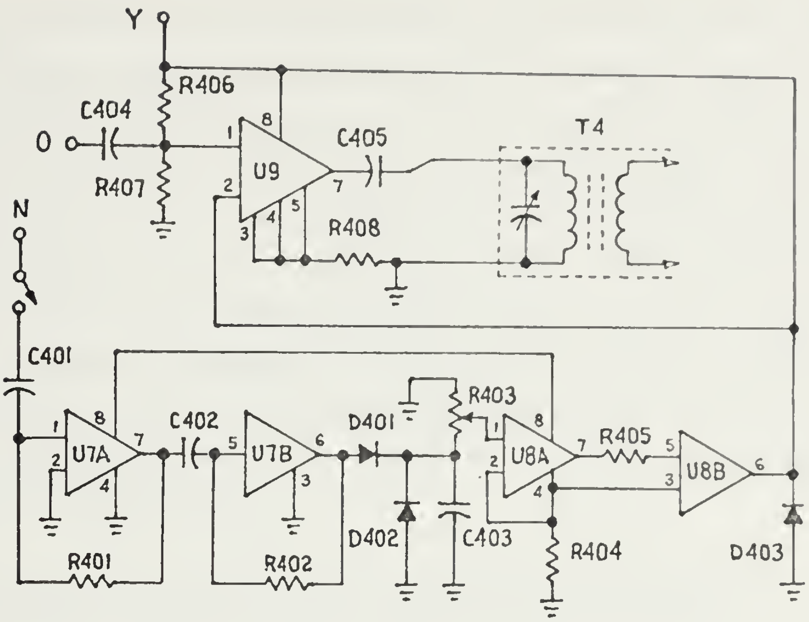

FIG. 29: MODIFIED NOISE BLANKER

The circuit shown in Fig. 29 is a Lamb silencer scheme offerred by Schulz [Ref. 6]. It uses three Fairchild uL914 integrated circuits, the circuit diagram for which is shown in Fig. 50 in Appendix D. The noise blanker has two inputs. Point N samples the mixer output. Point O connects to the output of the third IF amplifier. From the mixer, the IF signal is coupled to U7, which operates as a two stage common emitter amplifier. U8 is configured as a Schmidt trigger. The output of U8 is positive when the direct input voltage exceeds a certain value and is zero at other times. Positive pulses from U8 serve to cut off U9, which acts as a diode switch. R403 is the BLANKER control on the rear panel and sets the blanking level voltage above which the Schmidt trigger output is positive and U9 cuts off the IF signal. Switching is extremely fast in this circuit and high amplitude noise spikes should not be heard when

R403 is properly set.

This circuit was ultimately left out of the receiver operating signal line, although Fig. 3 show the intended noise blanker installation. Two circuit board errors survived initial checks and were discovered only after the circuit was tested. These remedies did not, however, solve the basic problem, which was a feedback route that caused oscillation in the first IF amplifier. Since the problem seemed to require a somewhat complex solution, such as a buffer between the mixer drain and point N in Fig. 29, the circuit was left disconnected while other more important problems were attacked.

4. Modified Receiver FM Detector and Squelch

Since the writer was interested in conducting future experiments with reception of VHF/UHF FM voice transmissions, an FM detector was included. A squelch amplifier was added to facilitate continuous monitoring of desired frequencies without unnecessary operator fatigue.

For FM signal detection, a Signetics N5111 integrated circuit was used in the circuit shown schematically in Fig. 30. This integrated circuit is functionally identical to the ULN2111A integrated circuit, also manufactured by Signetics. The FM detection technique of U10 features linear gating and is composed of a three stage limiting circuit, a balanced product detector, and associated biasing circuitry. This arrangement permits the use of a single inductor in the entire FM detector circuit. This inductor, T6, which is the high impedance winding of a 455 kHz transistor radio replacement transformer, is screwdriver adjusted to the receiver intermediate frequency. This circuit, used with success by Cross [Ref. 7], did not,

however, prove to be entirely compatible in parallel with the other detectors and, in the completed receiver, was disconnected from the signal line. The stage is included for reference purposes in appropriate diagrams and the schematic diagram of U10 is included as Fig. 51 in Appendix D.

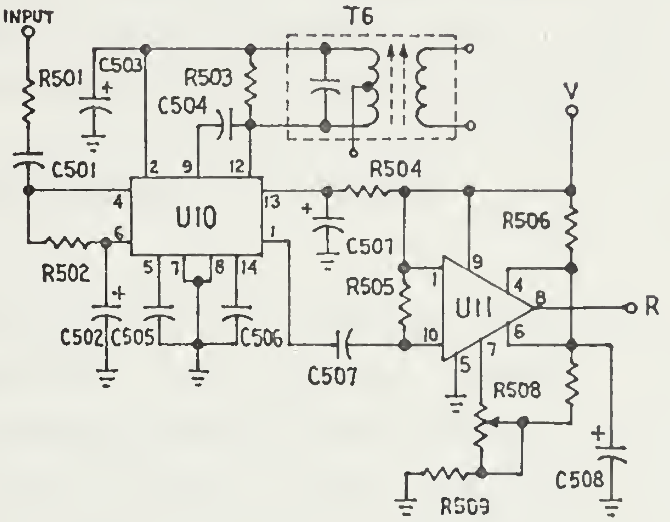

FIG. 30: FM DETECTOR & SQUELCH

A National LM370H was the integrated circuit used in the squelch circuit of Fig. 30, and was designated U11. U11 is a direct coupled monolithic amplifier whose gain is controlled by an external voltage applied to one of two control gates (pin 3 or pin 4.) These gates were not connected here, however, since this arrangement provides a fast attack, slow release squelch action independent of amplifier gain. Squelch threshold is set by potentiometer R508, mounted on the rear panel. A portion of the control voltage is injected at pin 6 of the integrated circuit to provide a hysteresis which reduces erratic response. Output of the squelch circuit leads from pin 8 of U11 to the mode

switch S5, on the receiver front panel. This output is designated point R in Fig. 30. The squelch circuit is a modified version of those recommended in the National applications manual [Ref. 8].

Since the FM detector stage never operated satisfactorily in the receiver circuit, the squelch circuit did not receive a truly fair evaluation of its performance. It did, however, appear to operate very well on signals injected at the input of U11 at various settings of R508.

5. Modified Receiver AM Detector

For AM reception, the infinite impedance detector of Fig. 31 was selected. This type of detector combines the high signal handling capabilities and low distortion of the diode detector and the high input impedance of the FET, which does not load the IF transformer to which it is connected. Q13 is an MPF102, which is operated as a source follower amplifier. The drain of Q13 is bypassed for both IF and AF while the source is bypassed only for IF. Thus IF signals are largely eliminated and the recovered audio is coupled through C513 to S5, the mode switch on the front panel. This circuit is described by Blakeslee [Ref. 2] and was used with minor modifications.

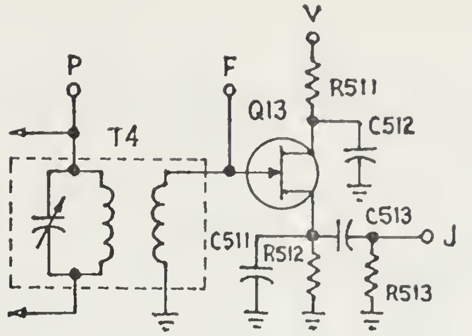

FIG. 32: MODIFIED AM DETECTOR

6. Modified Receiver Product Detector

The product detector circuit is shown in Fig. 32 and is described by Blakeslee [Ref. 2], Hejhall [Ref. 9], and the Motorcla integrated circuit data manual [Ref. 10]. It uses the Motorola MC1496G integrated circuit, an equivalent to identical devices offered by several other manufacturers. A circuit diagram of this device is shown as Fig. 53 in Appendix D. Included is a two transistor buffer, identical to those used in the IF amplifier stages, to match the low impedance integrated circuit input to the high impedance secondary of T4, which connects to point E. The second input, at point B, is the BFO injection. Output is at point S, which leads to the mode switch S5. Point V in Fig. 32 is the connection to the 12 volt supply line.

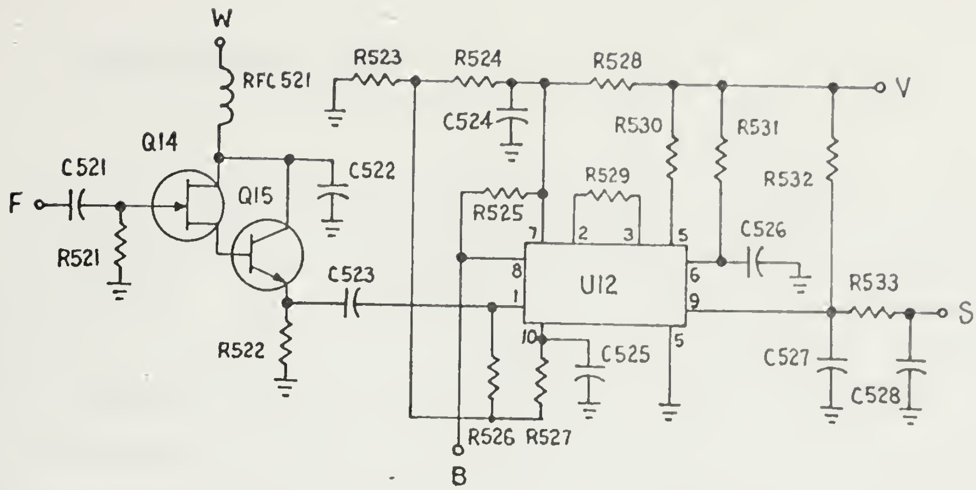

FIG. 32: MODIFIED PRODUCT DETECTOR

The principal advantage of using a product detector for enhanced reception of SSB and CW signals over the diode slope detector of the original SUPER-PRO is the substantial reduction of distortion and intermodulation products. In addition, the BFO signal need only be equal in amplitude to the input IF signal since the integrated circuit provides the requisite amplification for proper heterodyne operation. Output filtering is also simplified since the doubly balanced design of the MC1496G eliminates most of the IF and BFO components from the audio frequency output signal. Finally, the integrated circuit has a dynamic range of 90 dB and a conversion gain on the order of 12 dB, a characteristic which makes it a potent and distortionless audio driver stage. The product detector worked admirably once a good integrated circuit was obtained. The first device used had a base-to-emitter short in its BFO input transistor and therefore did not detect. A long delay in procuring a second device was compensated by the lack of any further trouble in the stage.

H. BEAT FREQUENCY OSCILLATOR STAGE

1. Original BFO Stage

The original BFO circuit, shown in Fig. 33, used a 6SJ7 pentode in an electron-coupled Hartley oscillator configuration. The oscillator coil L33 and associated components, mounted in a transformer enclosure similar to those containing the IF transformers, were together listed for convenience as T5. During alignment, C46 was adjusted to assure proper BFO operation at exactly 465 kilohertz. Variable capacitor C47 was the BEAT OSCILLATOR control accessible from the front panel and provided fine adjustment of the oscillator frequency during CW operation. Operating only when S3, the SIGNAL MOD-CW switch, was in the CW position, the BFO provided a 465 kilohertz signal to the primary of T4, where it was transformer coupled to the detector along with the IF signal. There, the two were mixed to allow detection of CW transmissions.

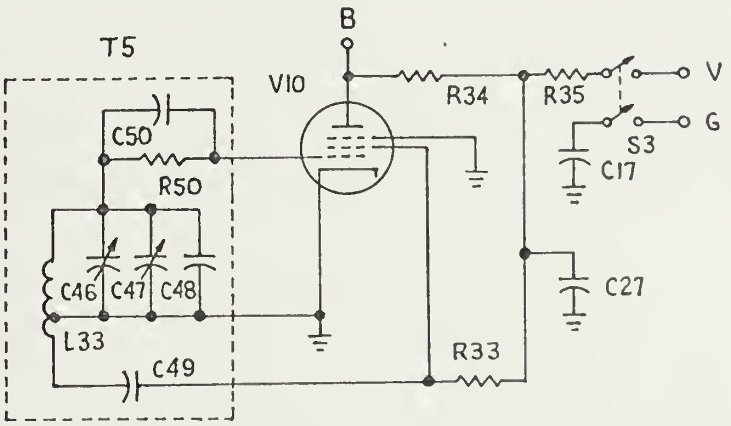

FIG. 33: ORIGINAL BFO

2. Modified BFO Module

The BFO module in the modified receiver, shown in Fig. 34, is actually composed of three oscillators and a two transistor buffer. Two of the oscillators are crystal-controlled to provide fixed reception of upper and lower sidebands. Transistors Q17 and Q18 are the oscillators for their respective sidebands and outputs are through C610 and C614 to the buffer. Q16 is the oscillator in the variable frequency portion of the module. This feature of the original SUPER-PRO was retained because of its particular usefulness in tuning CW signals. The circuit is a Seiler oscillator using a toroidal inductor in place of the original coil (L33 in Fig. 33,) which was accidentally ruined. Variable capacitors C46 and C47 were retained to perform their same functions in the new circuit. The output of the variable oscillator is coupled to the buffer via C606. The buffer is a copy of those used in the IF strip and product detector. Its output connects to point B of Fig. 32, the BFO input of the product detector. Point U is connected to the 9 volt supply line. The desired oscillator is selected by the mode switch S5B, which leads to the 12 volt line. In the AM and FM positions, all beat oscillators are inoperative. These oscillators and the buffer are described by Eslick [Ref. 5] and worked properly when built.

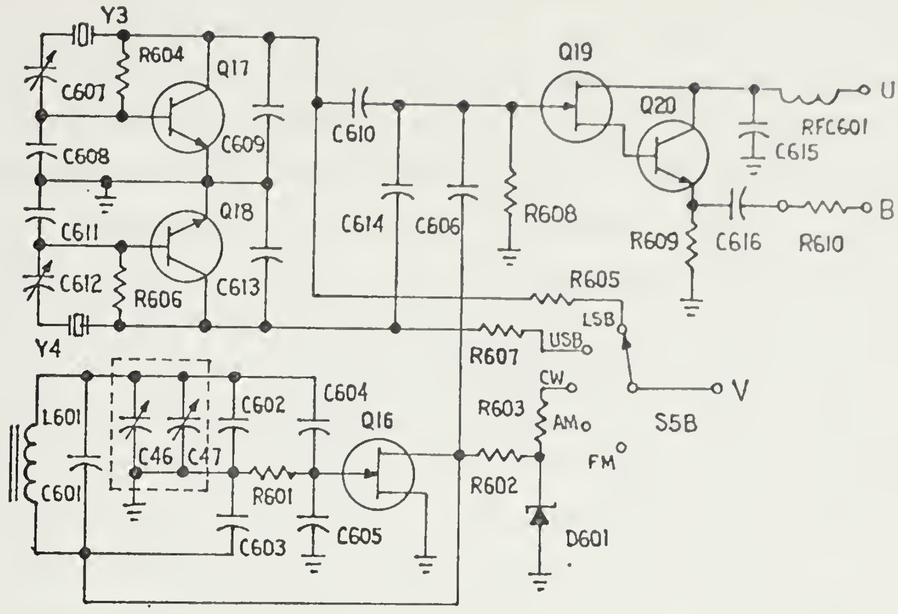

FIG. 34: MODIFIED BFO MODULE

I. AUTOMATIC VOLUME/GAIN CONTROL STAGES

1. Original AVC Stages

The original SUPER-PRO AVC circuit shown in Fig. 35 was composed of an AVC branch IF amplifier and an AVC rectifier-filter. The signal from the plate of the second IF amplifier tube V6 was coupled directly to the grid of V11, a 6SK7 pentode. Fixed grid bias was provided by the V7 grid circuit. This stage assured isolation between V6 and the AVC rectifier V12. V11 was coupled to V12 via transformer T6. V12 was a 6H6 dual diode with both sections wired in parallel and operated as a half-wave rectifier. Negative AVC voltage was developed across four resistors in series with V12 and was filtered by C54 and C55 to eliminate

ripple. AVC voltage to the RF amplifiers was coupled through S4, the AVC MANUAL-ON switch on the front panel. In addition, AVC voltage was coupled to the mixer and first two IF stages through potentiometer R56 and a resistive voltage divider. R56 was the SENSITIVITY control on the receiver front panel and also controlled the DC grid bias of V3, V5, and V6 when S4 was in the AVC ON position. When S4 was in the MANUAL position, no AVC voltage was provided and R56 controlled the DC grid bias of V1 and V2, as well as V3, V5, and V6. S meter operation occurred only when S4 was in the ON position. R41 was adjusted to read S9 on the meter scale for a 50 microvolt signal at 3.5 MHz and the BANDWIDTH control set to the 3 postion. AVC operation was quite rapid in the MOD position of S3, but was slowed in the CW position, the SIGNAL MOD-CW switch, by the longer time constant effected with the addition of capacitor C17 to the AVC line.

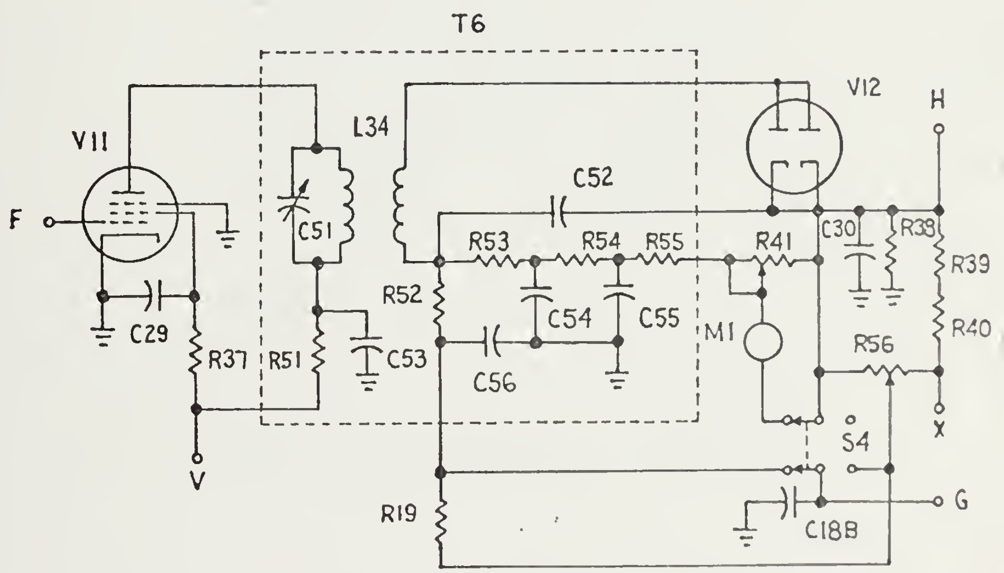

FIG. 35: ORIGINAL AVC SYSTEM

2. Modified AGC Stage

The new AGC stage is relatively simple, but surprisingly effective. It is built around the Plessey SL621C AGC generator integrated circuit, the schematic for which is shown in Fig. 54 in Appendix D. It possesses a fast attack, slow decay time constant required for optimum single-sideband reception. The integrated circuit input circuit activates a trigger providing a fast discharge path to the time constant capacitor for high level pulse signals, which would otherwise cause the circuit to suddenly increase the AGC voltage. In this way, the circuit nullifies the effect on the AGC of high amplitude noise spikes.

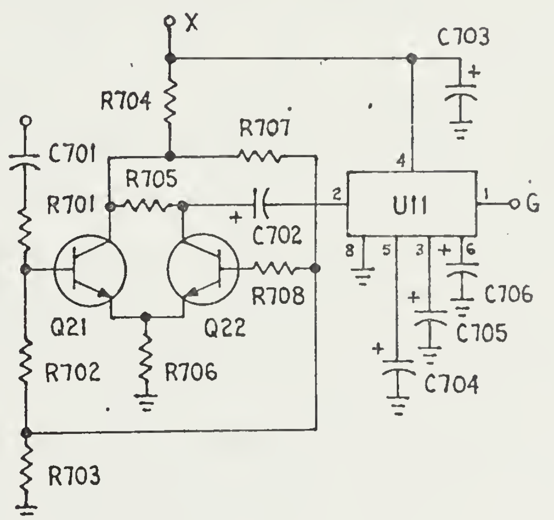

FIG. 36: MODIFIED AGC SYSTEM

The circuit ahead of U13, shown in Fig. 36, was recommended by Hrivnak [Ref. 11] for use with the SL621C. Q21 and Q22 essentially form an emitter coupled clipper and a resistive voltage divider. The former prevents the input

capacitor from charging while the latter is designed to keep the input signal level at 10 millivolts, as required for proper operation of U13. The remainder of the circuit is in accordance with the Plessey applications manual [Ref. 4] and uses only five electrolytic capacitors in addition to U13. Both sections operate on 6 volts drawn from the appropriate voltage supply at point X. Output at point G is fed to the first RF stage, the IF amplifiers, and the S meter amplifier.

3. Modified Receiver S Meter Amplifier

The S meter amplifier is a rather complex approach to the simple problem of measuring signal strength via the AGC voltage. In this circuit described by Blakeslee [Ref. 2] and shown in Fig. 37, Q23, an MPF102 field-effect transistor, samples the direct voltage of the AGC line at point G. The high input impedance of Q23 does not load the AGC generator. Q23 drives U14, an RCA CA3053 integrated circuit, whose relatively simple circuit diagram is shown in Fig. 55 in Appendix D. U14 provides the required current swing to operate the S meter, which is the only component of the original AVC system retained in this conversion. Its movement requires one milliampere for full scale deflection, as measured in the NPS Calibration Laboratory. This circuit is connected to the 12 volt supply line at point V in Fig. 37.

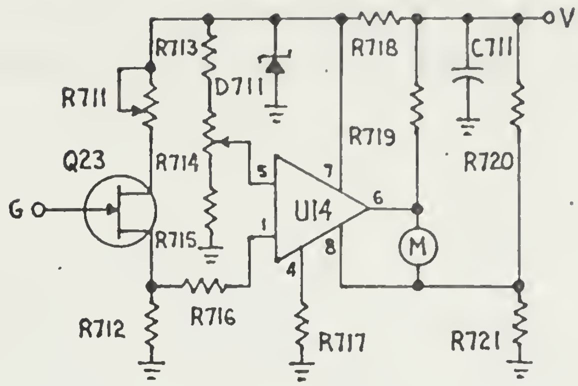

FIG. 37: MODIFIED S METER AMPLIFIER

J. AUDIO AMPLIFIER STAGES

1. Original Audio Amplifier Stages

The unmodified SUPER-PRO used a four tube audio amplifier stage, as shown in Fig. 38. Detector output was coupled to V13, the first AF amplifier, through the AUDIO GAIN control R26. V13 was a 6C5 triode which operated as a grounded cathode amplifier with output to the grid of V14. V14 was a 6F6 pentode, also operated in the grounded cathode configuration. Output of V14 went to the primary of T7, the push-pull input transformer. V15 and V16 were 6F6 pentodes operated as triodes in a push-pull audio final amplifier. The output of these tubes led to the primary of the push-pull output transformer T8, the center-tap of which connected to the +385 volt supply at point W. There were two secondary windings on T8 providing both 600 ohm speaker output and 8000 ohm headphone output. An auxiliary input to

the audio amplifier strip was available at point P. The amplifier provided up to 3 watts of output power with negligible distortion. Above 3 watts and up to a maximum output of 10 watts, grid current in V15 and V16 caused a corresponding increase in distortion of the audio output.

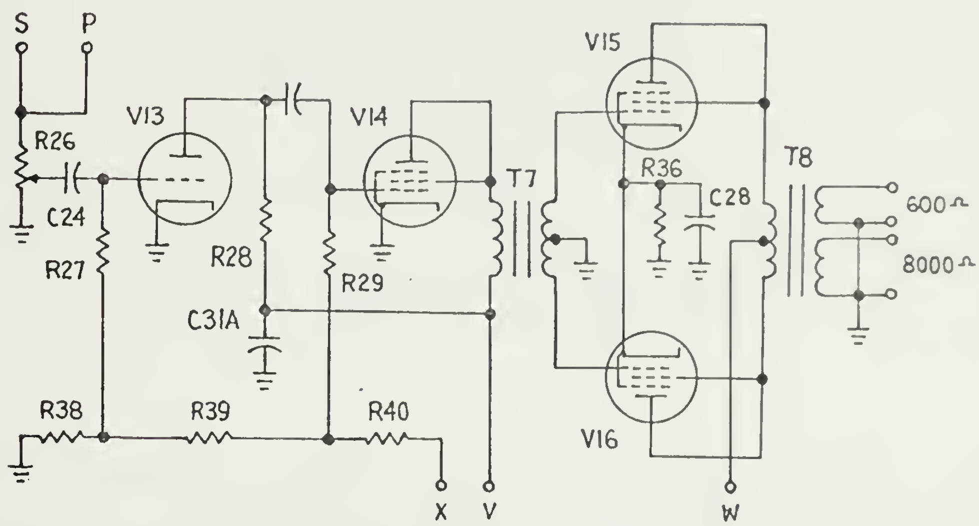

FIG. 38: ORIGINAL AUDIO AMPLIFIER

2. Interim Audio Amplifier

The interim audio amplifier was built to provide a simple and effective replacement for the four tube amplifier in the original circuit while other more critical stages were being worked on. Additionally, it permitted removal of the 385 volt line from the receiver. This line was used only by the push-pull output stage in Fig. 38 and represented a significant safety hazard while the equipment was energized.

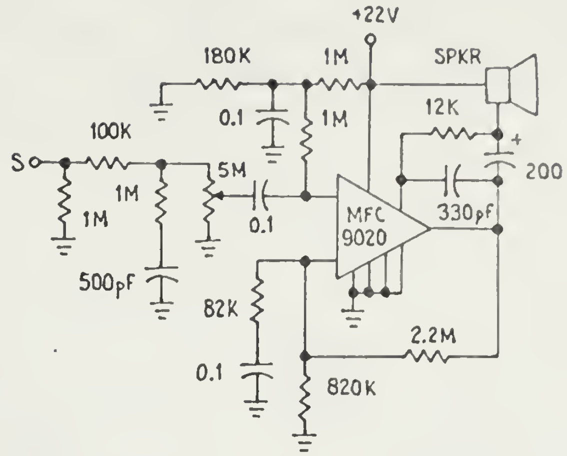

FIG. 39: INTERIM AUDIO AMPLIFIER

The Motorola MFC9020 integrated circuit used in the interim audio amplifier was capable of supplying two watts of continuous sine wave power with 0.2 volts RMS input, 22 volts DC supply, and minimal heat sinking. The circuit shown in Fig. 39 was borrowed from the manufacturer's data book with minor modifications. A 16 ohm speaker is specified, although an 8 ohm speaker performed adequately during construction and testing. A schematic of the MFC9020 is included in Appendix D as Fig. 59.

3. Modified Receiver Audio Module

Replacing the interim audio amplifier was a more sophisticated unit with the following features:

a. Active audio filter
b. Compressor
c. Preamplifier
d. Power amplifier

Designed by Megirian [Ref. 12], this module makes use of

three Signetics N5558V operational amplifiers. Each of these integrated circuits encloses two of the more familiar 741 type operational amplifiers in a single "mini-DIP" package. The last stage uses a Motorola MC1454G audio amplifier integrated circuit.

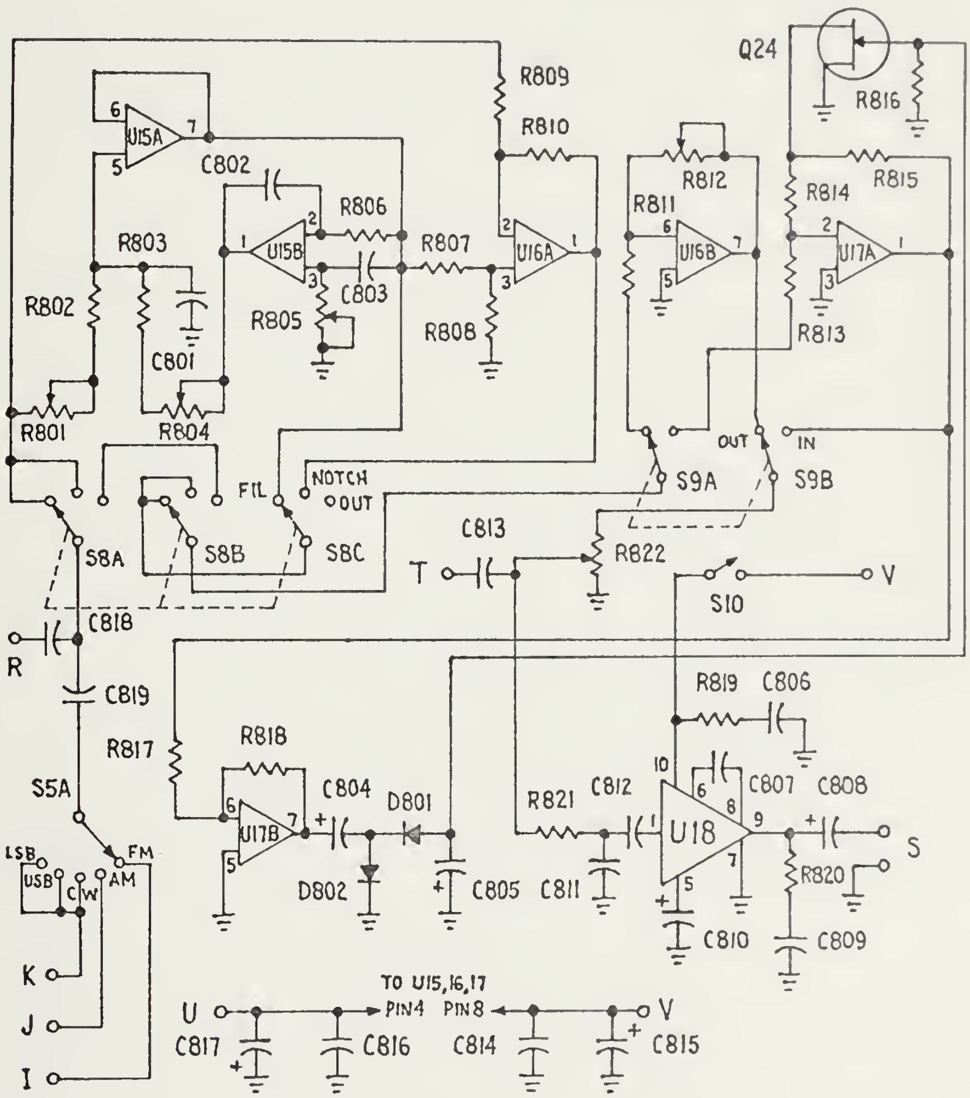

FIG. 40: MODIFIED AUDIO MODULE

The filter provides both bandpass and band reject functions with adjustable bandwidth and frequency. These controls (S8, R801, and R804 in Fig. 40) are mounted on the front panel. The filter is bypassed when not in use by means of the filter function switch S8. Figure 40 is the schematic diagram of the entire audio module. The center frequency of the filter is variable between approximately 300 Hertz and 2.0 kilohertz. The frequency is determined by an inductance simulating R-C network made up of C801, R803, and R804, the latter being a potentiometer and serving as the filter FREQUENCY control, and the operational amplifier U15B. Potentiometer R1 is the filter BANDWIDTH control. U16A is configured as a band reject filter in which equal and in-phase signals are applied to both inputs. At the notch frequency, the signals cancel, resulting in zero output. This feature is particularly useful in nulling heterodynes in AM reception. The sharpest filter frequency is about 11 Hertz at the -3 dB points of the filter response.

The compressor is actually an audio AGC system which provides flat output with low distortion from about .02 to 2.0 volts RMS input. The compressor may be switched out of the circuit by front panel control S9. This switch also inserts the preamplifier stage when the compressor is switched out to make up for the gain normally supplied by the compressor for small amplitude signals. Compression is accomplished by action of the N-channel FET Q24 in the feedback loop between the output of U17B and the input of U17A. Gain of the compressor at large signal amplitudes is limited by the maximum saturation resistance of Q24. AGC action has a fast attack, slow decay characteristic, the latter determined by the 50 microfarad value of capacitor C805.

The preamplifier, which operates when the compressor is not in use, is configured as a straight operational amplifier with gain adjustable by means of potentiometer R812 in Fig. 40.

The final audio amplifier integrated circuit has a one watt output capability. The circuit is taken from the Motorola integrated circuit data manual [Ref. 10]. A heat sink was provided for U18 to prevent overheating. Switch S10 turns off the power amplifier voltage supply for standby operation or use of high impedance headphones attached to point T in Fig. 40. The audio gain control R822 is adjusted from the front panel.

Power for the module is supplied from positive and negative polarity voltage lines at points V and U, respectively. The power amplifier also operates on the positive 12 volt supply line.

A minor circuit board design error prevented proper operation of this unit until it was discovered and corrected. Later, a bit of close wiring around the MODE switch S5 resulted in unwanted audio feedback and caused the amplifier to "motorboat" at audio gain settings above midscale. Use of shielded wire around the critical area cleared up the problem.

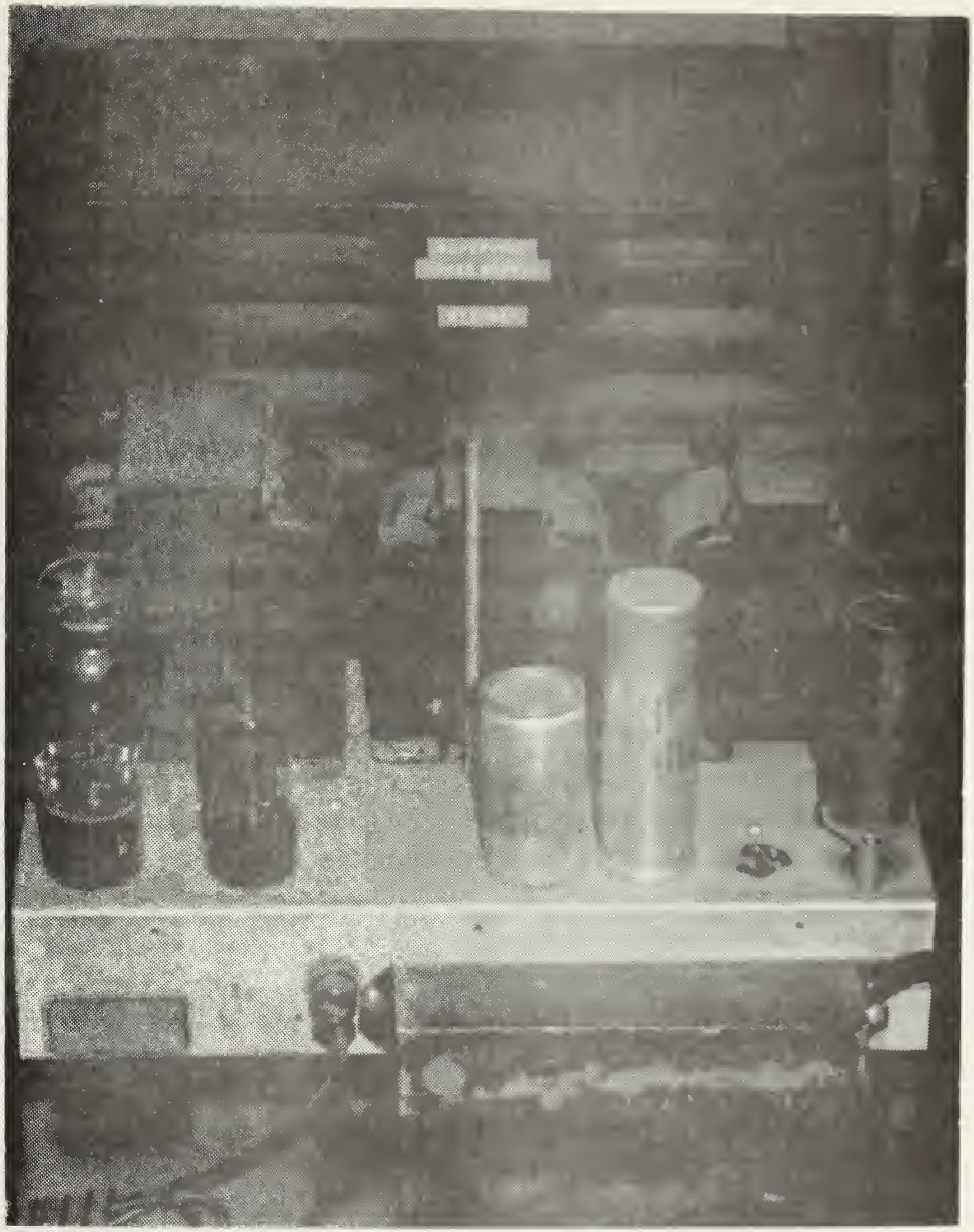

FIG. 41: ORIGINAL SUPER-PRO POWER SUPPLY

K. POWER SUPPLIES

1. Original Power Supply

The original SUPER-PRO power supply was built on a separate chassis as shown in the photograph of Fig. 41. The schematic diagram of this unit is shown in Fig. 42. There was nothing remarkable about the power supply, except that a separate rectifier and filter system was used to provide negative grid bias voltage. The single power transformer had three primary taps for various source voltages and a tap on the high voltage secondary to provide low voltage for the bias supply. The high voltage rectifier tube was a type 5Z3 and the bias rectifier was a type 80. The high voltage circuit used a two stage capacitive input filter. The 385 volt line, however, was drawn from between the two filter chokes to avoid the significant voltage drop across L2, whose inductance was valued at 50 Henries. A voltage divider circuit, composed of L2, R1A, and R1B provided the lesser supply voltages. The bias supply used a three stage capacitive input filter with resistors in place of inductors as the series elements of the filter. Each rectifier tube was provided with its own filament voltage winding and a third such winding supplied the receiver 6.3 volt AC filament and dial lamp potential.

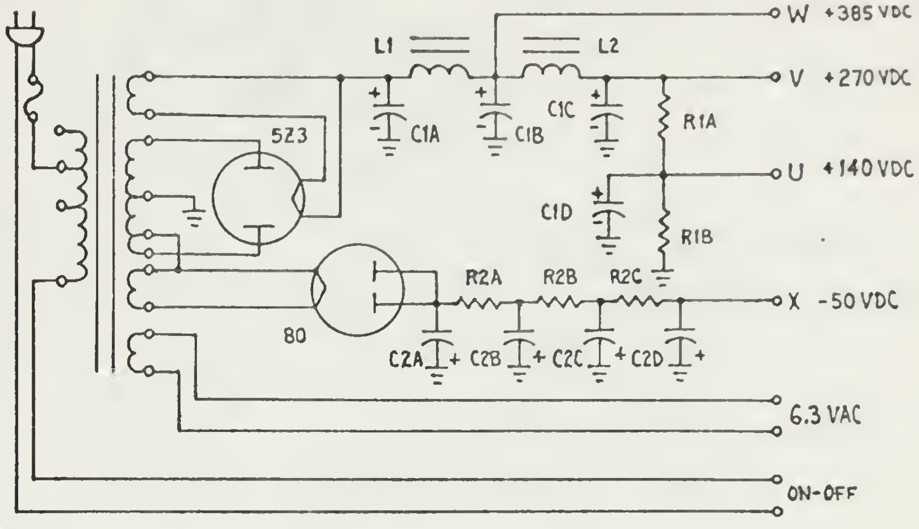

FIG. 42: ORIGINAL POWER SUPPLY

2. Modified Receiver Power Supply

The modified SUPER-PRO power supply was entirely solid-state and was built on an etched circuit board that is mounted on the receiver chassis, as shown at the upper left in Fig. 7. The circuit diagram of this power supply is shown in Fig. 43. There are three secondary windings on the power transformer T7. The uppermost winding provides potential to incandescent lamps I1, I2, and I3, which illuminate the bandspread dial, main tuning dial, and S meter, respectively. This voltage is rectified and capacitively filtered to eliminate possible 60 Hertz pickup in the nearby audio gain control leads. Average direct voltage is approximately 4.6 volts, which gives satisfactory illumination and should extend average bulb life by almost an order of magnitude.

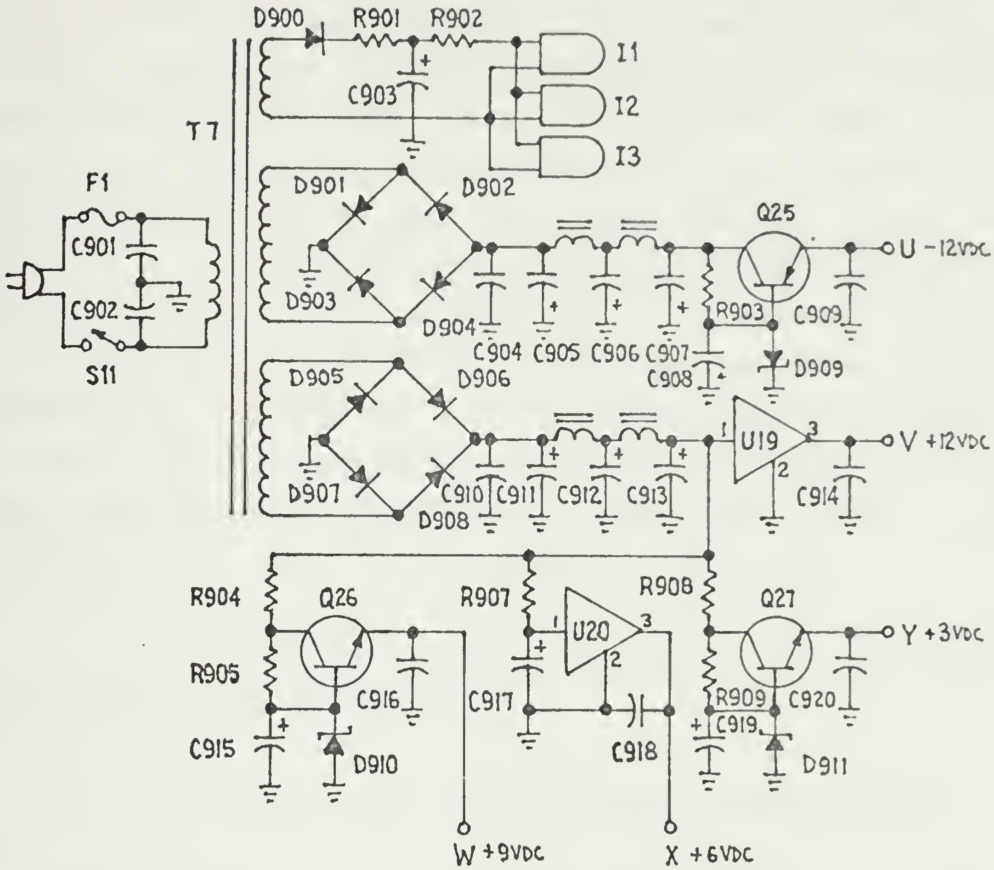

FIG. 43: MODIFIED POWER SUPPLY

The central secondary of T7, as shown in Fig. 43, feeds the negative supply furnishing voltage to the audio module. A bridge rectifier preceeds the two stage capacitive input filter composed of electrolytic capacitors C905, C906, and C907 and the 100 millihenry chokes L901 and L902 (unnumbered in Fig. 43.) A series regulator assures operation at a constant voltage of approximately 11.8 volts.

The lower winding voltage is rectified and filtered by a full-wave bridge and filter circuit similar to the system described above, except for reversed polarities on

all diodes and electrolytic capacitors. Four voltage regulating circuits draw their inputs from this source. The +12 and +6 volt supplies make use of Motorola MC7800 series voltage regulator integrated circuits, whose employment was described by Math [Ref. 13], and Trulove [Ref. 14]. The units are designed to deliver up to 15 watts of power without use of an external pass transistor. They are also internally protected against excess output current, output short circuit, and overheating. A circuit diagram for integrated circuits of this series is shown in Fig. 58 in Appendix D. In the +12 volt circuit, U19 is an MC7812P and in the +6 volt supply, U20 is an MC7806P. The +9 and +3 volt supplies use simple series regulator similar to that described for the -12 volt supply above.

Current drain on the respective voltage supplies is tabulated below in Table II.

TABLE II. POWER SUPPLY CURRENT DRAIN

Supply Voltage	Typical Current Drain
-12 V.	8 ma.
+12 V.	126 ma.
+9 V.	89 ma.
+6 V.	23 ma.
+3 V.	52 ma.

III. ANALYSIS OF RESULTS

A. PERFORMANCE COMPARISON

The project was nominally completed when the performance of the modified SUPER-PRO approximated that of the original receiver. In several respects, the solid-state version is superior to that of the vacuum tube model. Other characteristics were inferior to those of the original SUPER-PRO.

Since only one frequency band was investigated in detail, a sensitivity comparison is limited to the 2.5 to 5.0 MHz. band. The sensitivity of the modified receiver at the low and high frequency ends of the band was measured in the same manner as with the original receiver and Table III shows the results compared with those listed in Table I for the band specified.

TABLE III. SENSITIVITY COMPARISON

Frequency (MHz.)	Original (uV.)	Modified (uV.)
2.5	16.0	1.9
5.0	5.9	2.0

Clearly, the modified receiver is superior in terms of sensitivity. The probable reason for this disparity is the gain achieved in the second RF amplifier, the entire IF amplifier strip, and the product detector. So much gain, in

fact, is developed that tightening the variable coupling of the coils in the adjustable IF transformers T2 and T3 results in oscillation in the IF strip. A large amount of gain, however, is needed to compensate for the inability to properly tune the inductors of the front end tuning assembly and the primaries of the IF transformers. This phenomenon reduces the response of these tuned circuits to a point generally on the high frequency side of the resonance point. Lending weight to this suspicion is the fact that, while tuning through a CW signal, a much greater response is measured on the low frequency side of zero-beat than on the high frequency side.

As previously described, a persistent and rapid frequency drift characterized the original SUPER-PRO. This drift was observed by tuning the receiver to zero-beat with the carrier of WWV with the SIGNAL MOD-CW switch set to CW. In the modified receiver, a digital frequency counter was coupled to the local oscillator output at gate two of the mixer transistor Q4 with the receiver dial set to 5.0 MHz. A continuous read-out was obtained and the frequency drift with respect to time after power was switched on is presented in Table IV below.

TABLE IV. MODIFIED SUPER-PRO FREQUENCY DRIFT

Time (min.)	Drift (kHz.)
0	0.0
1	-0.1
10	-0.3
30	-0.6
100	-1.0

This drift characteristic, while not favorably comparable to modern receivers, is considerably better than that of the original SUPER-PRO, whose drift was both rapid and continuous. Since both the original and modified receivers drifted toward lower frequencies, it is apparent that common components may be involved. Since tuned circuitry in the tuning box was electrically unchanged in the modification, the cause of the drift is probably contained therein. Aged capacitors, temperature-affected inductors, or dirty bandswitch contacts are likely culprits in the drift situation.

Mechanical stability of the modified receiver is vastly improved over the original version. As previously mentioned, the SUPER-PRO was most sensitive to physical shock and vibration. Investigation showed that the wiring in the tuning box used light (22 gauge) insulated hook-up wire in lengths long enough to vibrate visibly at the slightest jar of the receiver. This obviously contributed to the mechanical instability of the receiver and stouter wiring was clearly needed. Enamelled 14 gauge wire was available and was a good fit for the solder lugs on the tuned circuit components. At considerable effort, all tuned circuit wiring was replaced by the heavier wire. Although

impact on the receiver chassis still causes a frequency jump, the shift is usually momentary and a much harder blow is needed to cause it. In addition, the drop of a pencil or similar object in the vicinity of the receiver no longer causes the frequency to waver. Clearly, the heavier wire effected a much-needed improvement

The modified receiver has serious problems with intermodulation products, or "birdies." Both the local oscillator/Mixer and the BFO module are mounted atop the chassis and neither is isolated by a shielding box, as is the IF amplifier strip. Non-sinusoidal waveforms are produced by the two crystal-controlled beat oscillators and the harmonics present when either of these oscillators is operating undoubtedly contributes to the intermodulation problem. No effort was made to clear up the trouble, although shielding of the oscillators would be a first approach.

Other aspects of the receiver conversion that differed between the original and modified versions are briefly as follows: Modified receiver AM reception suffers because of the lack of gain following the detector stage, as previously discussed. The power consumption of the modified SUPER-PRO is only 12 watts as compared to 180 watts for the original receiver and power supply. A substantial weight reduction was effected in the modification. The modified receiver weighed 44 pounds, as compared to a 83 pounds for the original receiver and external power supply combination.

B. PROJECT FEASIBILITY

The project was admittedly ambitious. The large number of tuned circuits to be rematched to the differing input and output impedances and capacitances, while maintaining the original performance characterisitcs, presented formidable problems, not all of which were solved. The use of field-effect transistors in the first RF amplifier, mixer, local oscillator, and AM detector stages partially solved the replacement problems, although the input capacitance of the FET is generally much greater than that of the pentode. Since the use of integrated circuits was of primary interest in this project, they were used in several stages where FETs would perhaps have been better suited. To make up for the mismatches inherent in the use of integrated circuits in place of vacuum tubes, there was liberal use of buffer stages with emitter/source follower configurations. The stages following the IF amplifier strip were, however, more ideally suited for replacement with integrated circuits. As a result, fewer problems and better results were experienced in these stages. An alternative to the problems of impedance and capacitance matching in the RF and IF stages would have involved discarding existing tuned circuits and using specially constructed new ones. This, however, would have been tantamount to building a new receiver on the old chassis and all real identity of the SUPER-PRO design would have been lost. The project was feasible as conceived, but less than spectacular results were anticipated. The foremost goal in the project was to gain practical experience in electronics and that goal was achieved. The modified receiver operates in a satisfactory manner and offers promise for interesting and valuable independent research in the future.

C. EXPERIENCE ASPECTS OF THE PROJECT

In the fourteen months in which the project was completed, much was learned by the writer in terms of practical electronics. No attempt is made to list all facets of this experience, but the most significant aspects are described below.

Perhaps the most important benefit of this project is the everyday familiarity with a wide variety of electronic test equipment. The specific test equipment used in this project is listed in Appendix E. Regular laboratory experiments in the course of instruction at NPS provided a valuable introduction to many types of test equipment. This introduction was, however, insufficient, in the writer's opinion, to provide a familiarity with test equipment expected in a modern electrical engineer. The writer indeed feels fortunate that this project has afforded him that kind of experience. Further, it has provided to him the incentive to obtain a modest set of quality instruments for future experiments in electronics.

A second valuable lesson concerned component tolerance and reliability. Widely varying characteristics were noted in the transistors, both field-effect and junction types, used in the solid-state conversion. The semiconductor curve tracing oscilloscope proved to be an invaluable asset in choosing transistors for optimum performance, particularly in the AM detector and audio compressor stages. A few components (the product detector and some 0.1 microfarad bypass capacitors,) were defective and the improbability of this circumstance effectively taught the author valuable lessons in trouble-shooting.

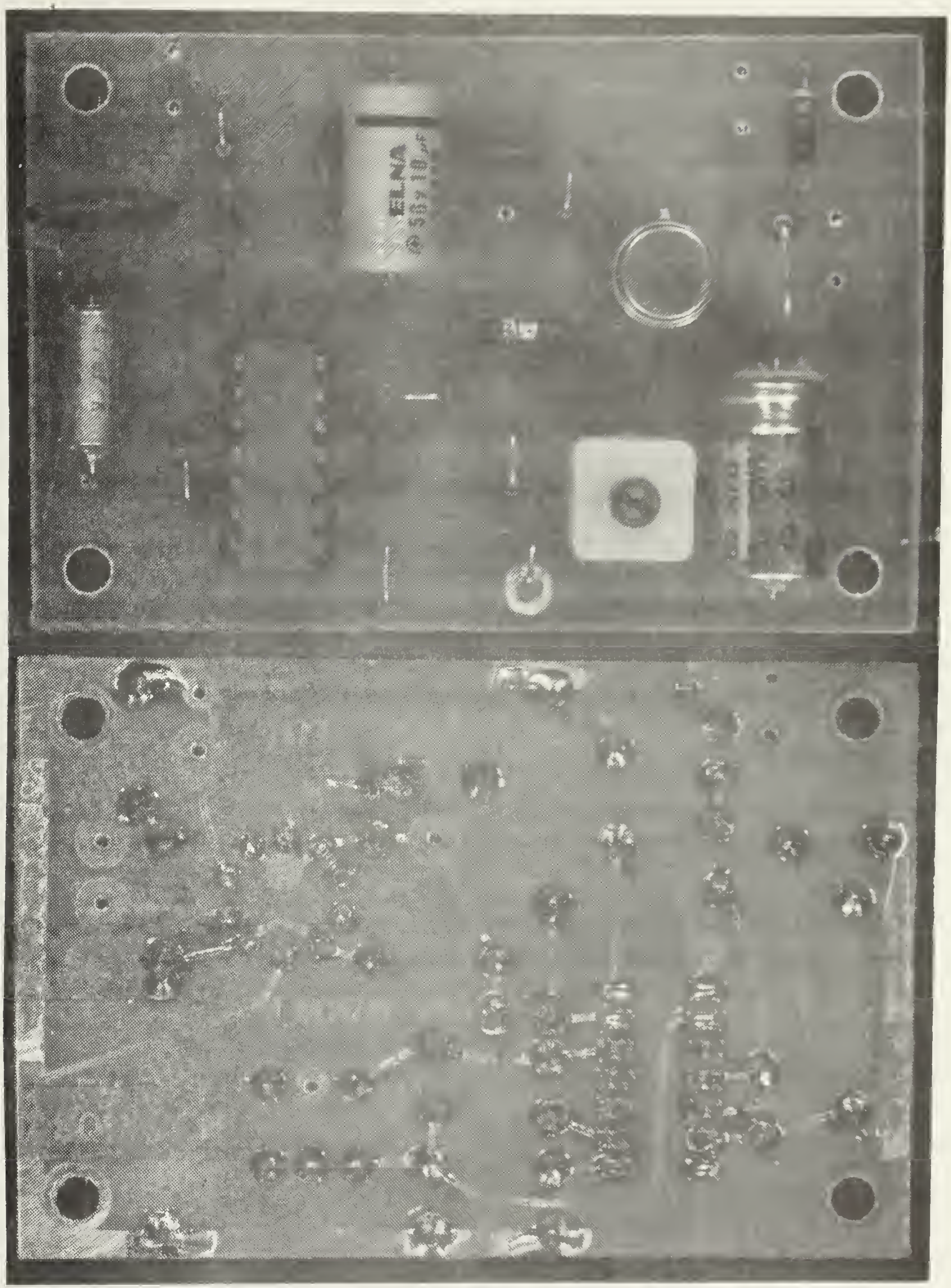

FIG. 44: SAMPLE ETCHED CIRCUIT BOARD
(TOP: COMPONENT SIDE, BOTTOM: FOIL SIDE)

Full use was made of the NPS Etching Laboratory and considerable improvement was noted in the writer's design of etched circuit boards as the conversion progressed. Only eleven circuit boards are mounted on the completed receiver chassis, but several others were discarded for poor design or replaced by boards for redesigned circuits. An understanding of the fabrication of etched circuit boards was gained by soliciting instruction from the laboratory technicians.

Interstage impedance matching is a very important consideration which was dealt with many times during the course of the project, not always with complete success. The use of emitter/source follower buffer stages to match impedances was thoroughly investigated.

A real understanding of Q was finally gained in the course of the project. Consideration of the inductors in the front end and the problems associated with the possibilities of replacing the low frequency coverage with 20 to 30 MHz coverage led to a study of Q and some experiments with a Q meter. The idea of extended high frequency coverage was ultimately abandoned, but its consideration was nonetheless very instructive.

The original receiver, as previously mentioned, drifted severely in frequency. Several measures were used in an attempt to reduce or eliminate the drift. The most effective measure proved to be the application of non-residue spray cleaner to the contacts of the bandswitch. This measure was recommended by the Hammarlund Company in response to inquiry on the subject. The drift was not entirely eliminated, but the study of the problem was

certainly worthwhile.

Many other useful lessons were learned such as the cause and cure of "motorboating" in the audio stage, IF and RF alignment procedure, crystal oscillator operation, use of toroidal inductors, regulated power supply operation and construction, transformer construction, and the use of semiconductor data and application manuals. This kind of experience should be very helpful in any future work with electronics.

D. FUTURE POSSIBILITIES

As previously mentioned, the writer intends to continue work on this project in the future as time and service obligations permit. Many improvements and changes are contemplated. Planned initial modifications are listed below as follows:

a. Trial replacement of the present IF strip by a second strip using dual-gate MOSFETs.

b. Installation of an integrated circuit preamplifier following the AM detector.

c. Installation of a buffer between T4 and the FM detector.

d. Elimination or reduction of intermodulation problems.

e. Construction of a replacement noise blanker stage using dual-gate MOSFETs.

f. Installation of input trimming circuit in the first RF amplifier stage as recommended by Knietel [Ref. 1].

g. Modification of front end tuned circuits for the 5.0 - 10.0 MHz. and 10.0 - 20.0 MHz. bands.

Ultimately the writer plans to replace the existing front end with a voltage tuned system with digital frequency read-out. Crystal-controlled converters for HF and partial VHF coverage would then be added. The large size of the receiver chassis and the large area to be made available by removal of the massive RF tuning box facilitiate future modifications.

IV. CONCLUSIONS

The replacement of vacuum tubes by semiconductor devices in a relatively sophisticated communications receiver is a feasible project in terms of material results. Dramatic improvement of receiver performance characteristics is not to be expected without a major redesign of the receiver and a more realistic project would be the design and construction of an entirely new receiver to suit specified requirements. In terms of practical experience in electronics, the re-engineering of a vacuum tube receiver to solid-state operation is, however, an excellent project for the serious student of electronics, supplementing his practical experience gained in scheduled laboratory exercises and complementing his theoretical knowledge of electronics learned in the classroom.

APPENDIX A BC-779A SPECIFICATIONS AND TUBE LIST

SPECIFICATIONS:

FREQUENCY COVERAGE: 100-200 kilohertz
200-400 kilohertz
2.5-5.0 megahertz
5.0-10.0 megahertz
10.0-20.0 megahertz

RECEIVER TYPE: Superheterodyne

NUMBER OF TUBES: 18

INTERMEDIATE FREQUENCY: 465 kilohertz

TYPES OF RECEPTION: CW and AM

POWER INPUT: 180 watts

RECEIVER WEIGHT: 55 pounds

POWER SUPPLY WEIGHT: 28 pounds

TUBE LIST

BC-779A RECEIVER:

V1	6K7	Variable-mu pentode
V2	6K7	Variable-mu pentode
V3	6L7	Pentagrid mixer
V4	6J7	Sharp cut-off pentode
V5	6K7	Variable-mu pentode
V6	6SK7	Variable-mu pentode
V7	6SK7	Variable-mu pentode
V8	6H6	Dual diode
V9	6N7	Class B dual triode

V10	6SJ7	Sharp cut-off pentode
V11	6SK7	Variable-mu pentode
V12	6H6	Dual diode
V13	6C5	Medium-mu triode
V14	6F6	Power amplifier pentode
V15	6F6	Power amplifier pentode
V16	6F6	Power amplifier pentode

POWER SUPPLY:

V1	5Z3	Full wave rectifier
V2	80	Full wave rectifier

APPENDIX B BC-779A SCHEMATIC DIAGRAM

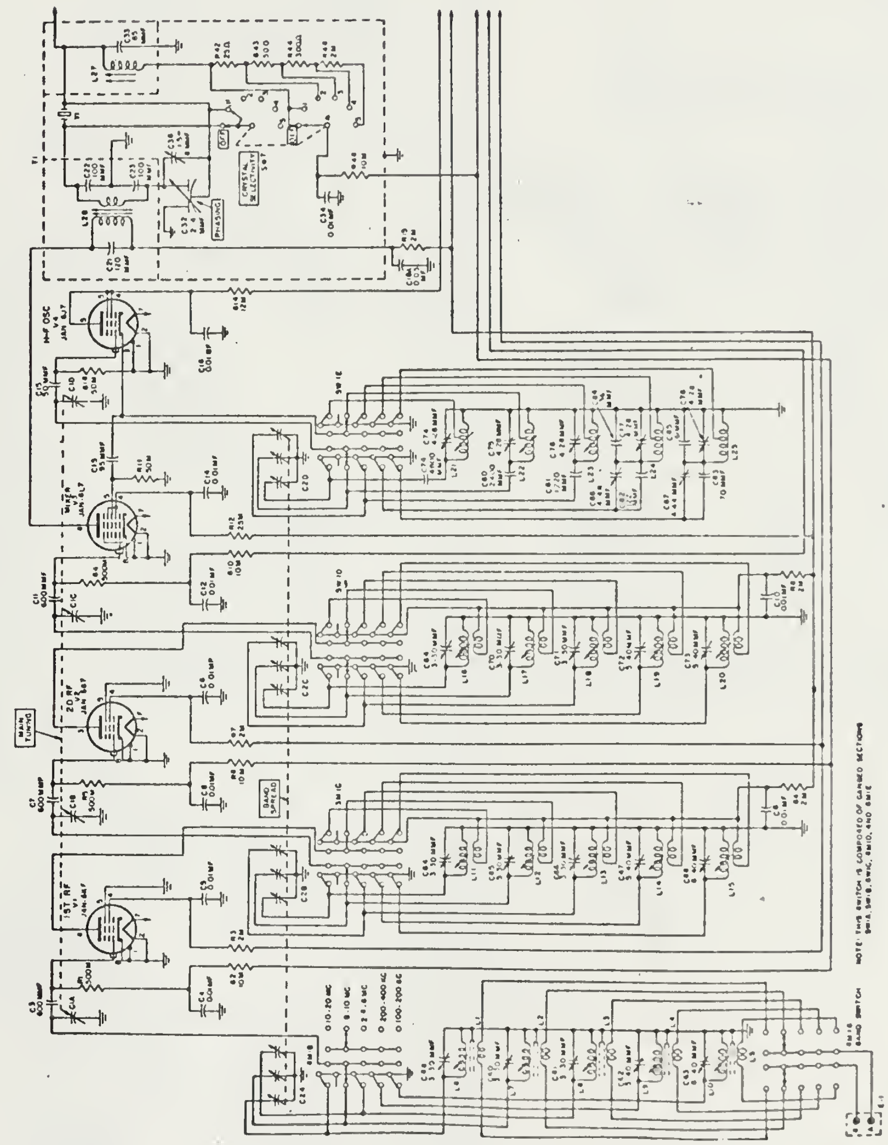

FIG. 45A: BC-779A SCHEMATIC DIAGRAM

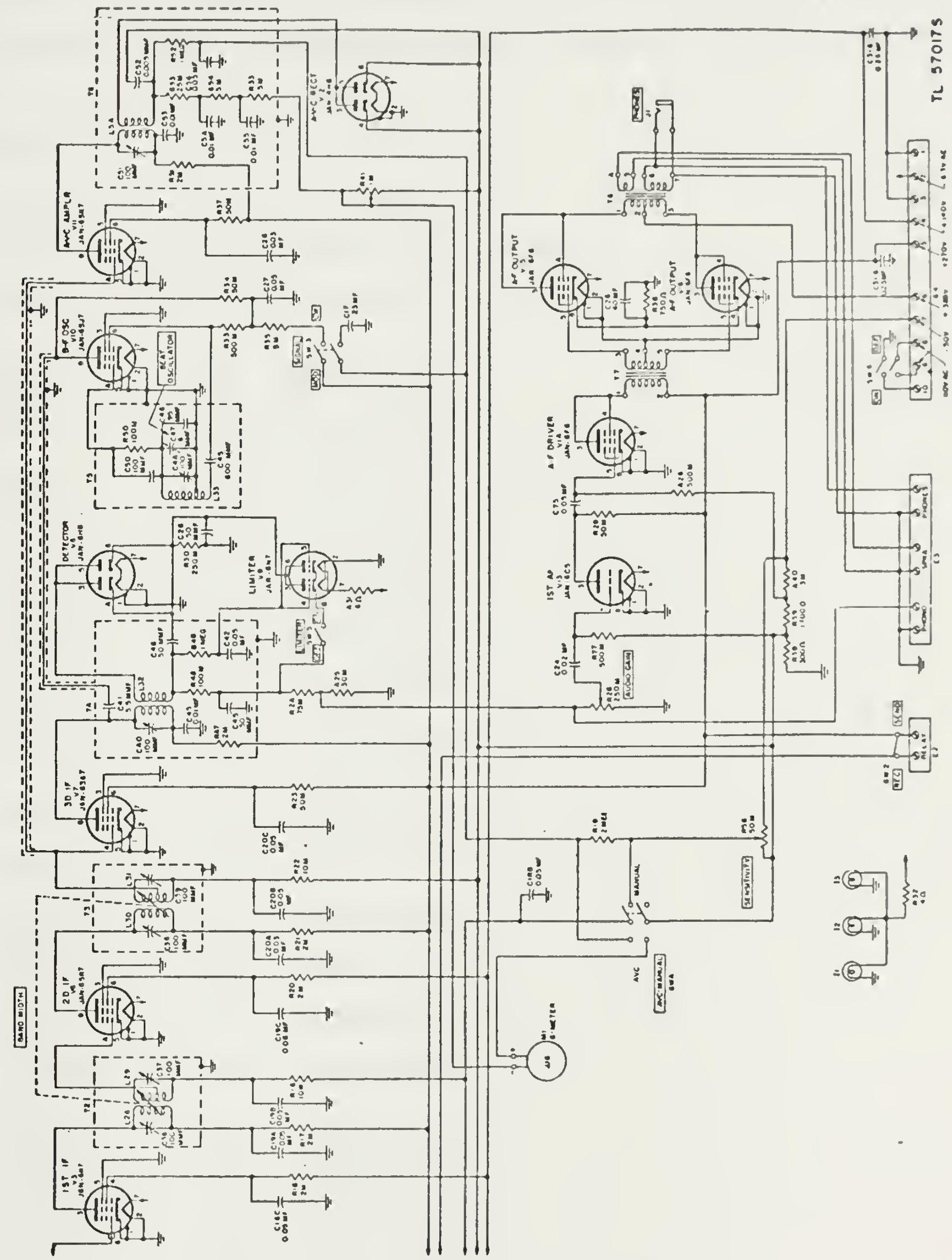

FIG. 45B: BC-779A SCHEMATIC DIAGRAM

APPENDIX C SOLID-STATE SUPER-PRO COMPONENT LIST

Below are listed components added during modification with nominal values indicated and special details noted.

CAPACITORS

C001	.1	microfarad
C002	2.5-30	picofarad trimmer
C003	24	picofarad
C101	270	picofarad
C102	.01	microfarad
C103	.01	microfarad
C111	.015	microfarad
C112	.01	microfarad
C113	.015	microfarad
C114	.01	microfarad
C115	.01	microfarad
C201	.01	microfarad
C202	.01	microfarad
C211	.05	microfarad
C212	220	picofarad
C213	.01	microfarad
C214	24	picofarad
C301	.15	microfarad
C302	.15	microfarad
C303	.15	microfarad
C304	.15	microfarad
C305	.15	microfarad
C311	.15	microfarad
C312	.15	microfarad

C313	.15	microfarad
C314	.15	microfarad
C315	.15	microfarad
C321	.15	microfarad
C322	.15	microfarad
C323	.15	microfarad
C324	.15	microfarad
C325	.15	microfarad
C401	100	picofarad
C402	.001	microfarad
C403	.001	microfarad
C404	.01	microfarad
C405	.001	microfarad
C501	.005	microfarad
C502	5	microfarad electrolytic
C503	5	microfarad electrolytic
C504	68	picofarad
C505	.1	microfarad
C506	.005	microfarad
C507	.02	microfarad
C508	50	microfarad electrolytic
C511	500	picofarad
C512	.5	microfarad
C513	.1	microfarad
C521	.15	microfarad
C522	.15	microfarad
C523	.15	microfarad
C524	.1	microfarad
C525	.1	microfarad
C526	.005	microfarad
C527	.005	microfarad
C528	.005	microfarad
C601	33	picofarad ceramic NPO

C602	270	picofarad
C603	2200	picofarad
C604	510	picofarad
C605	2200	picofarad
C606	500	picofarad
C607	100	picofarad trimmer
C608	1500	picofarad
C609	1500	picofarad
C610	1000	picofarad
C611	1500	picofarad
C612	100	picofarad trimmer
C613	1500	picofarad
C614	1000	picofarad
C615	.15	microfarad
C616	.022	microfarad
C701	4.7	microfarad electrolytic
C702	5	microfarad electrolytic
C703	50	microfarad electrolytic
C704	50	microfarad electrolytic
C705	100	microfarad electrolytic
C706	100	microfarad electrolytic
C711	.02	microfarad
C801	.15	microfarad
C802	.005	microfarad
C803	.005	microfarad
C804	5	microfarad electrolytic
C805	50	microfarad electrolytic
C806	.1	microfarad
C807	39	picofarad
C808	100	microfarad electrolytic
C809	.1	microfarad
C810	10	microfarad electrolytic
C811	.005	picofarad
C812	1	microfarad electrolytic

C813	.47	microfarad
C814	.1	microfarad
C815	50	picofarad electrolytic
C816	.1	microfarad
C817	50	microfarad electrolytic
C818	.1	microfarad
C819	1	microfarad electrolytic
C901	.01	microfarad
C902	.01	microfarad
C903	470	microfarad electrolytic
C904	.01	microfarad
C905	200	microfarad electrolytic
C906	470	microfarad electrolytic
C907	470	microfarad electrolytic
C908		microfarad electrolytic
C909	.1	microfarad
C910	.01	microfarad
C911	200	microfarad electrolytic
C912	470	microfarad electrolytic
C913	470	microfarad electrolytic
C914	.1	microfarad
C915	470	microfarad electrolytic
C916	.1	microfarad
C917	470	microfarad electrolytic
C918	.1	microfarad
C919	200	microfarad electrolytic
C920	.1	microfarad

RESISTORS

R001	110	kilohm
R002	110	kilohm
R101	330	ohm
R101	1	megohm
R103	1	megohm

R104	100	ohm
R111	100	kilohm
R112	10	kilohm
R113	4.7	kilohm
R114	470	ohm
R115	47	ohm
R201	100	kilohm
R202	270	ohm
R203	150	ohm
R211	47	kilohm
R212	100	kilohm
R213	100	ohm
R214	560	ohm
R301	110	kilohm
R302	1	kilohm
R303	10	kilohm
R304	1	kilohm
R305	100	ohm
R311	110	kilohm
R312	1	kilohm
R313	10	kilohm
R314	1	kilohm
R315	100	ohm
R321	110	kilohm
R322	1	kilohm
R323	10	kilohm
R324	1	kilohm
R325	100	ohm
R401	33	kilohm
R402	33	kilohm
R403	2.5	kilohm potentiometer
R404	47	ohm

R405	910	ohm
R406	10	kilohm
R407	11	kilohm
R408	220	ohm
R501	22	ohm
R502	470	ohm
R503	22	kilohm
R504	100	ohm
R505	1	kilohm
R506	110	kilohm
R507	33	kilohm
R508	10	kilohm potentiometer
R509	200	ohm
R511	6.8	kilohm
R512	27	kilohm
R513	47	kilohm
R521	110	kilohm
R522	100	ohm
R523	1	kilohm
R524	820	ohm
R525	47	ohm
R526	4.7	kilohm
R527	4.7	kilohm
R528	1.2	kilohm
R529	100	ohm
R530	10	kilohm
R531	2.7	kilohm
R532	2.7	kilohm
R533	1	kilohm
R601	47	kilohm
R602	150	ohm
R603	10	ohm
R604	110	kilohm
R605	680	ohm

R606	110	kilohm
R607	680	ohm
R608	110	kilohm
R609	100	ohm
R610	47	ohm
R701	15	kilohm
R702	10	kilohm
R703	6.8	kilohm
R704	270	ohm
R705	100	ohm
R706	3.3	kilohm
R707	4.7	kilohm
R708	10	kilohm
R711	10	kilohm potentiometer
R712	4.7	kilohm
R713	15	kilohm
R714	10	kilohm potentiometer
R715	4.7	kilohm
R716	4.7	kilohm
R717	100	ohm
R718	100	ohm
R719	220	ohm
R720	220	ohm
R721	1	megohm
R801	100	kilohm potentiometer
R802	560	ohm
R803	1	kilohm
R804	50	kilohm potentiometer
R805	10	kilohm potentiometer
R806	6.8	kilohm
R807	10	kilohm
R808	10	kilohm
R809	10	kilohm
R810	10	kilohm

R811	2	kilohm
R812	100	kilohm potentiometer
R813	100	kilohm
R814	10	kilohm
R815	10	kilohm
R816	100	kilohm
R817	10	kilohm
R818	56	kilohm
R819	10	ohm
R820	10	ohm
R821	1	kilohm
R822	10	kilohm potentiometer
R901	0	(removed)
R902	0	(removed)
R903	1	kilohm
R904	22	ohm
R905	1	kilohm
R907	110	ohm
R908	110	ohm
R909	1	kilohm

DIODES

D101	1N100	
D102	1N100	
D211	1N914	
D401	1N914	
D402	1N914	
D403	1N914	
D601	1N757	9.0V. zener
D711	1N757	9.0V. zener
D801	1N914	
D802	1N914	
D900	1N538	
D901	1N3612	
D902	1N3612	

D903	1N3612	
D904	1N3612	
D905	1N3612	
D906	1N3612	
D907	1N3612	
D908	1N3612	
D909	1N759A	12.6V zener
D910	1N3011	9.0V zener
D911	HEPZ0208	3.9V zener

TRANSISTORS

Q1	HEP802	field-effect
Q2	HEP802	field-effect
Q3	2N3705	
Q4	MPF122	MOS field-effect
Q5	MPF102	field-effect
Q6	40673	MOS field-effect
Q7	MPF102	field-effect
Q8	2N3706	
Q9	MPF102	field-effect
Q10	2N3706	
q11	MPF102	field-effect
Q12	2N3706	
Q13	MPF102	field-effect
Q14	MPF102	field-effect
Q15	2N3706	
Q16	MPF102	field-effect
Q17	2N3706	
Q18	2N3706	
Q19	MPF102	field-effect
Q20	2N3706	
Q21	2N3904	
Q22	2N3904	
Q23	MPF102	field-effect
Q24	HEP801	field-effect

Q25	2N1041
Q26	TIP33A
Q27	2N3866

CRYSTALS

Y1	465.00 kilohertz (crystal filter)
Y2	100.00 kilohertz (marker generator)
Y3	463.35 kilohertz (BFO - USB)
Y4	466.65 kilohertz (BFO - LSB)

RADIO FREQUENCY CHOKES

RFC211	2.5 millihenries
RFC301	10 millihenries
RFC302	10 millihenries
RFC303	10 millihenries
RFC521	10 millihenries
RFC601	10 millihenries

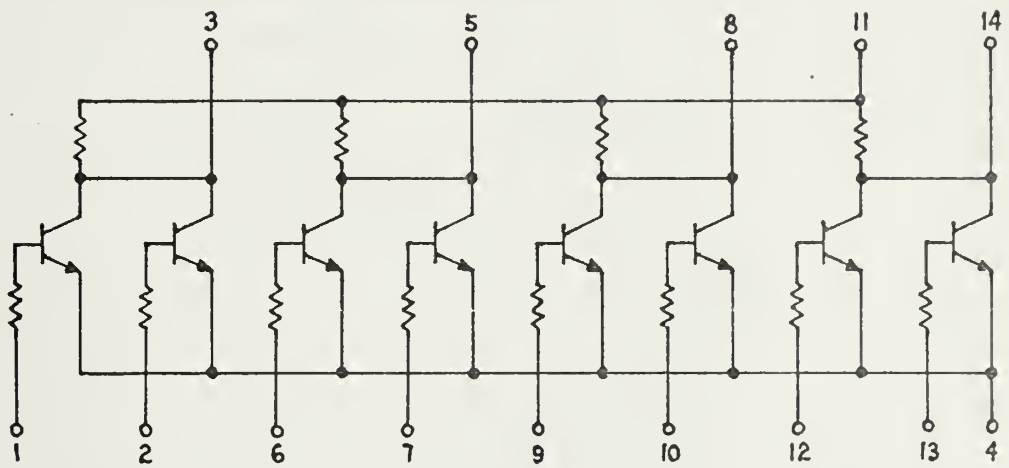

FIG. 46: MOTOROLA MC724P (U1)

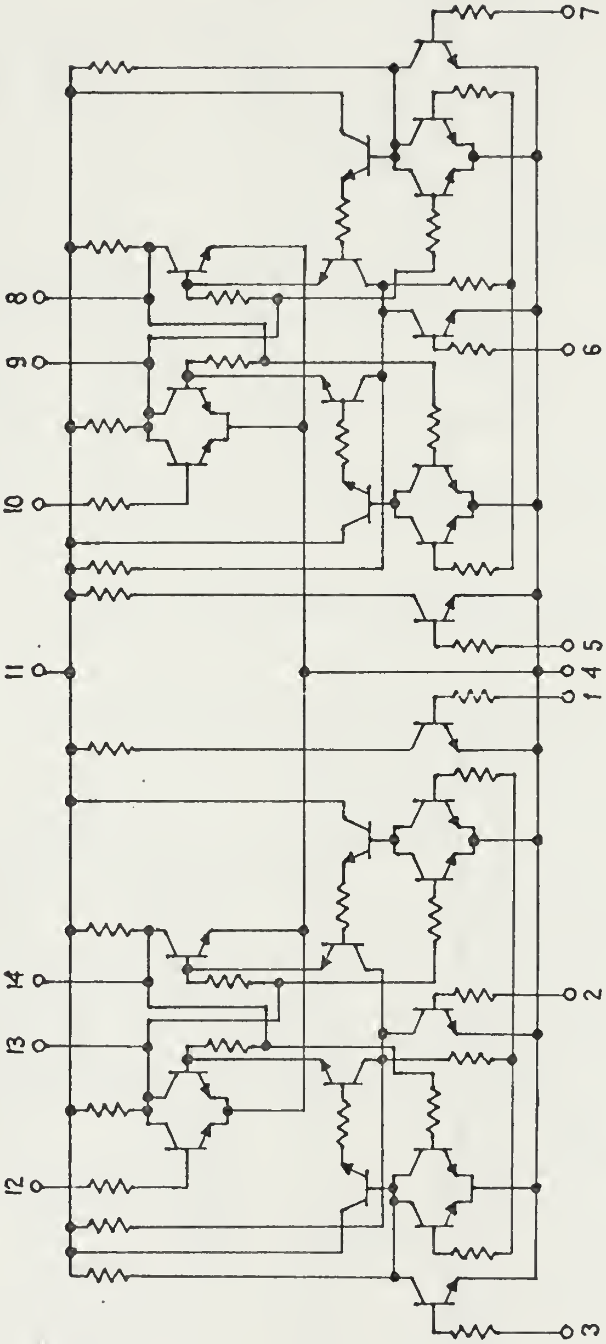

FIG. 47: MOTOROLA MC790P (U2)

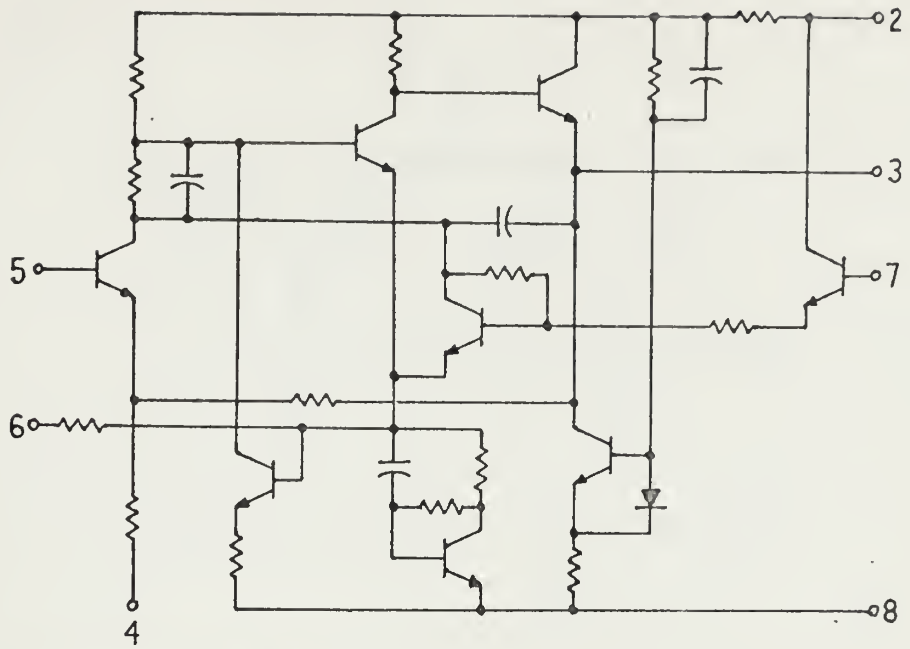

FIG. 48: PLESSEY SL610C (U3)

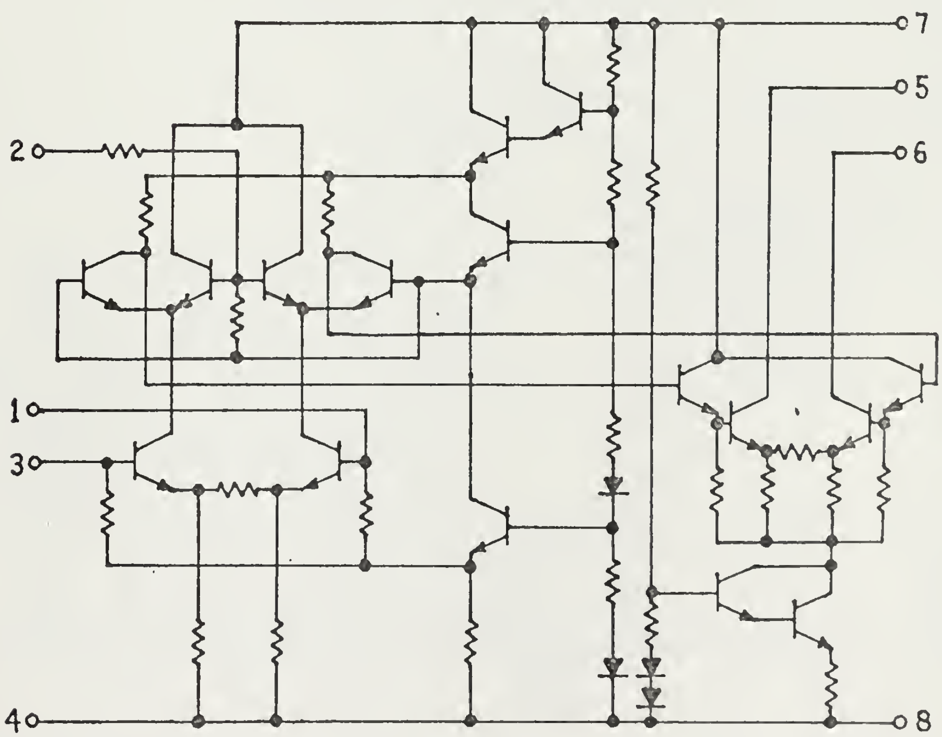

FIG. 49: MOTOROLA MC1590G (U4,5,6)

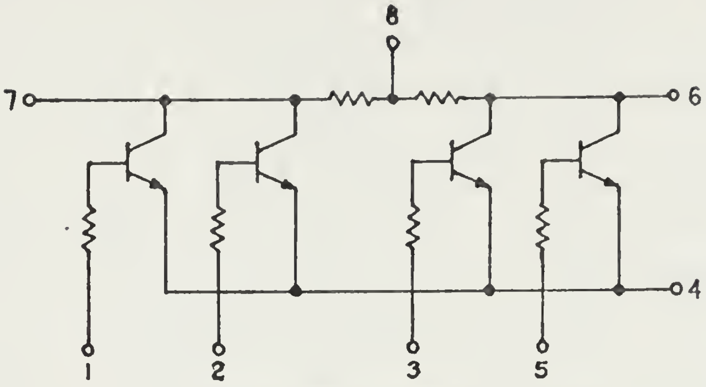

FIG. 50: FAIRCHILD UL914 (U7,8,9)

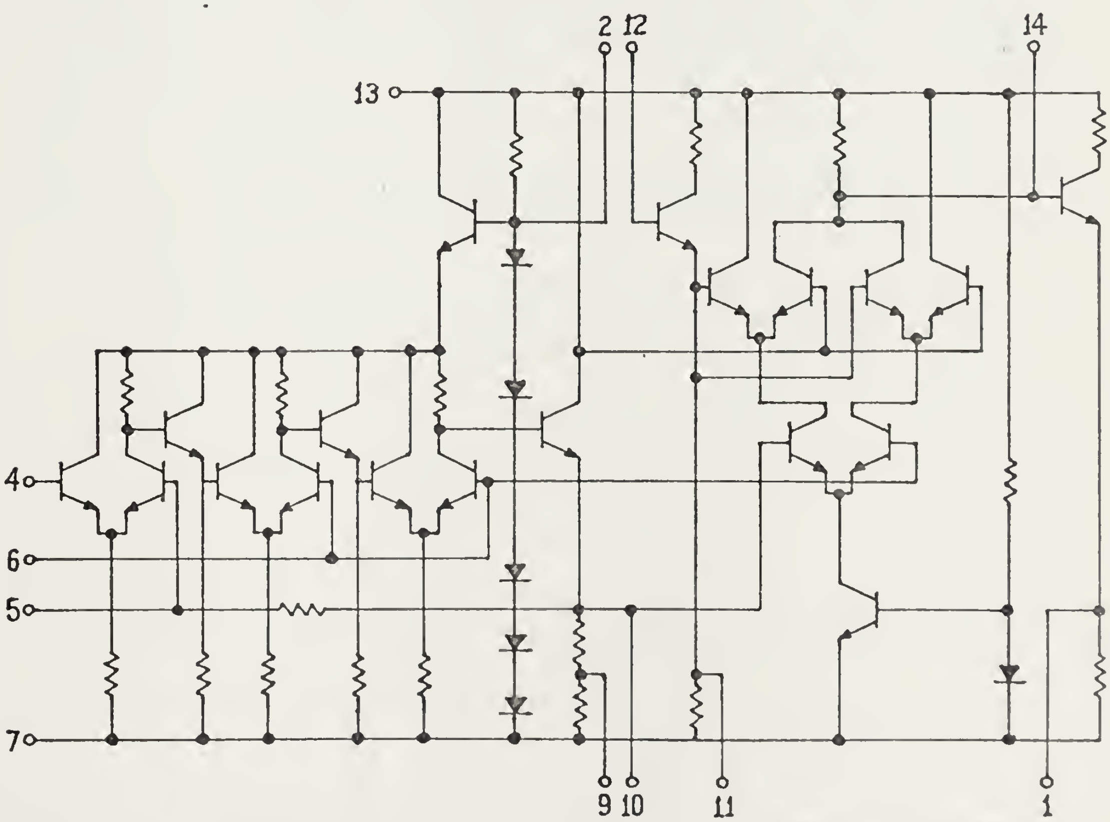

FIG. 51: SIGNETICS N5111 (U10)

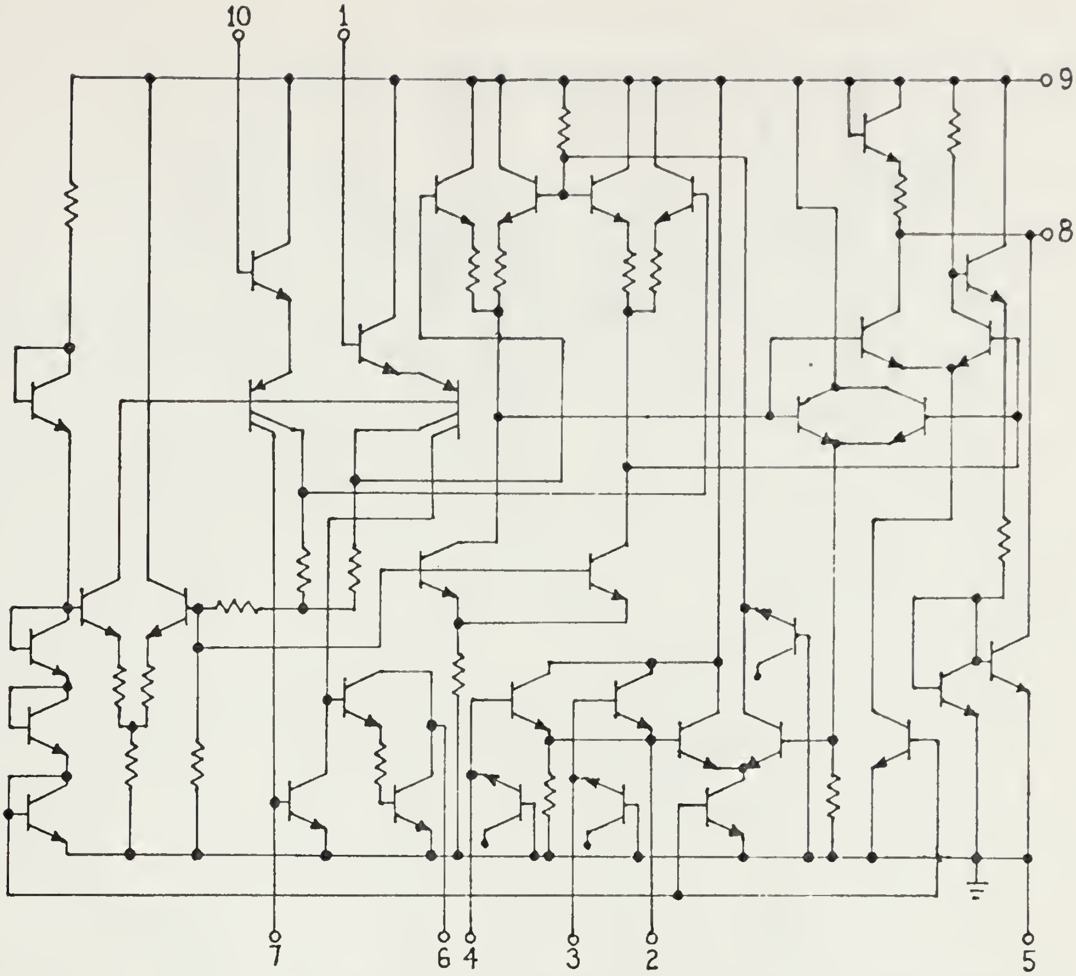

FIG. 52: NATIONAL LM370H (U11)

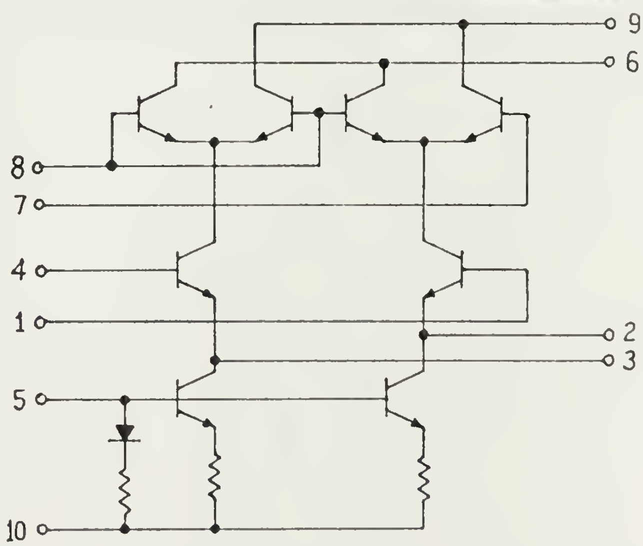

FIG. 53: MOTOROLA MC1496G (U12)

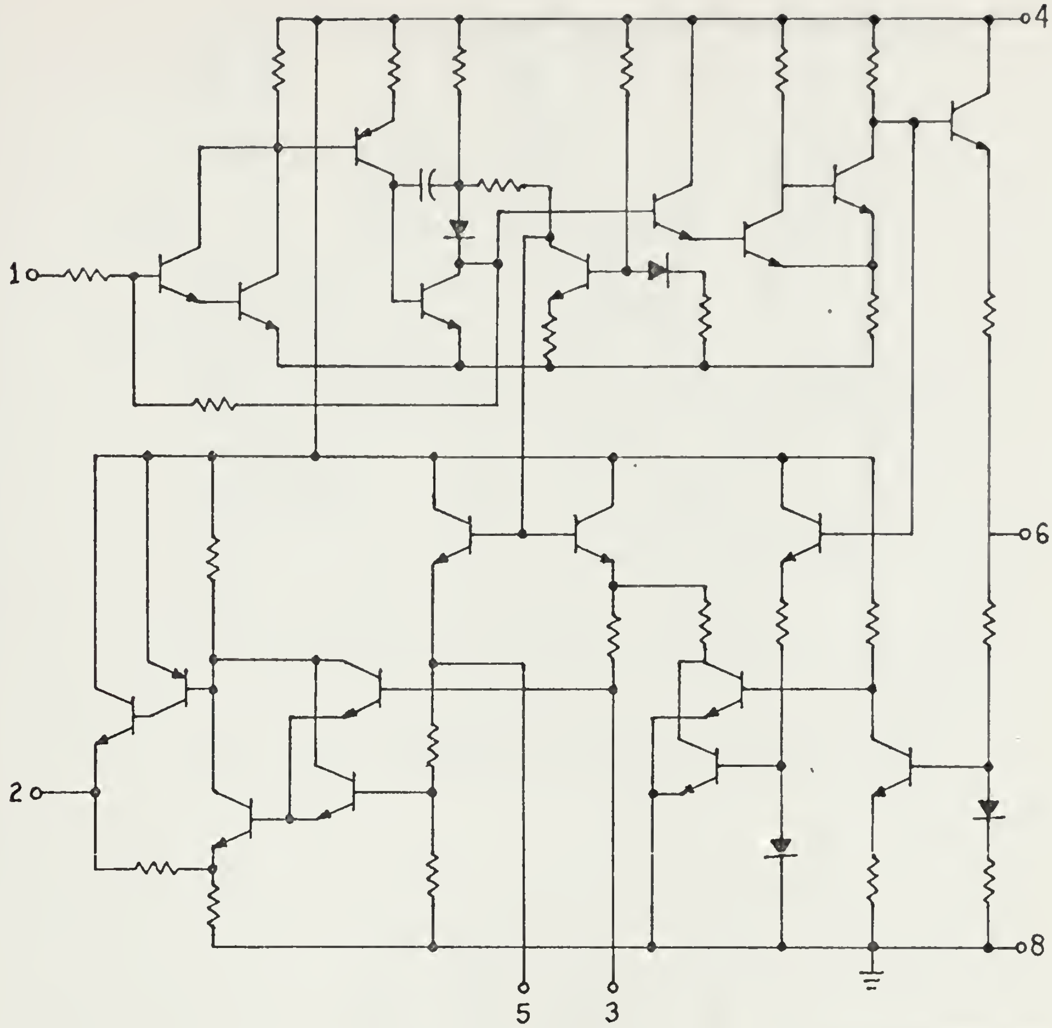

FIG. 54: PLESSEY SL621C (U13)

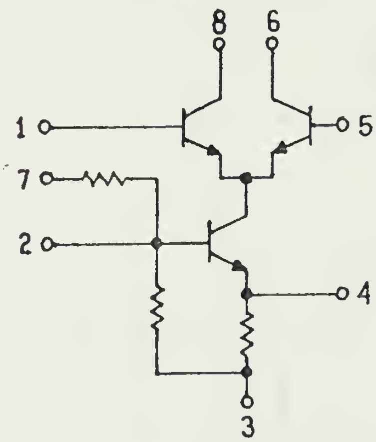

FIG. 55: RCA CA3053 (U14)

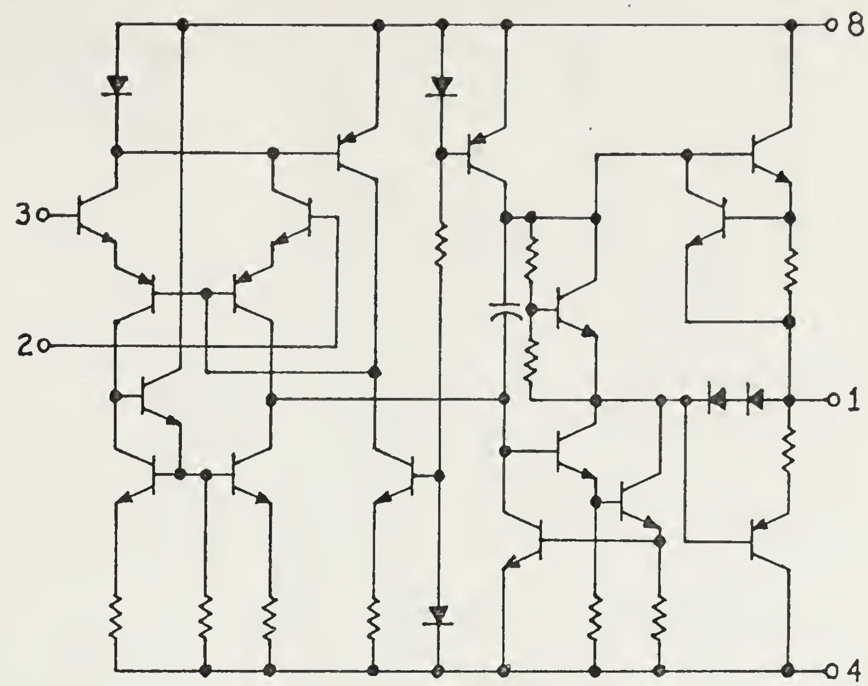

FIG. 56: SIGNETICS N5558V (U15,16,17)

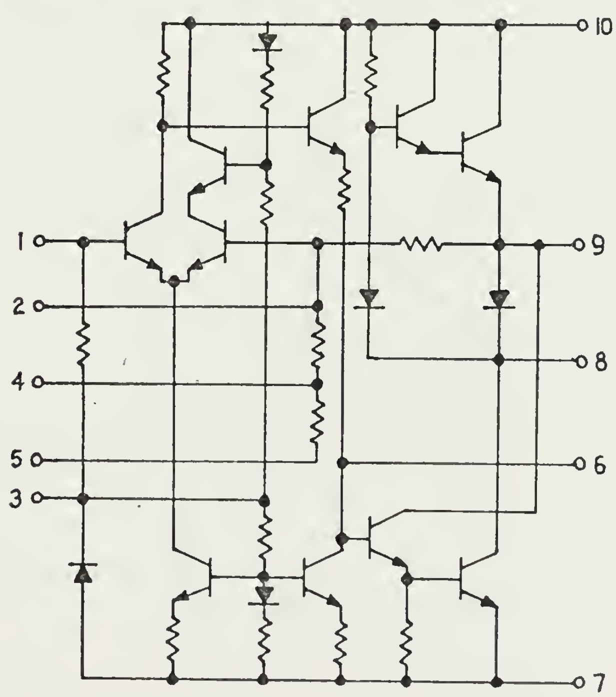

FIG. 57: MOTOROLA MC1454G (U18)

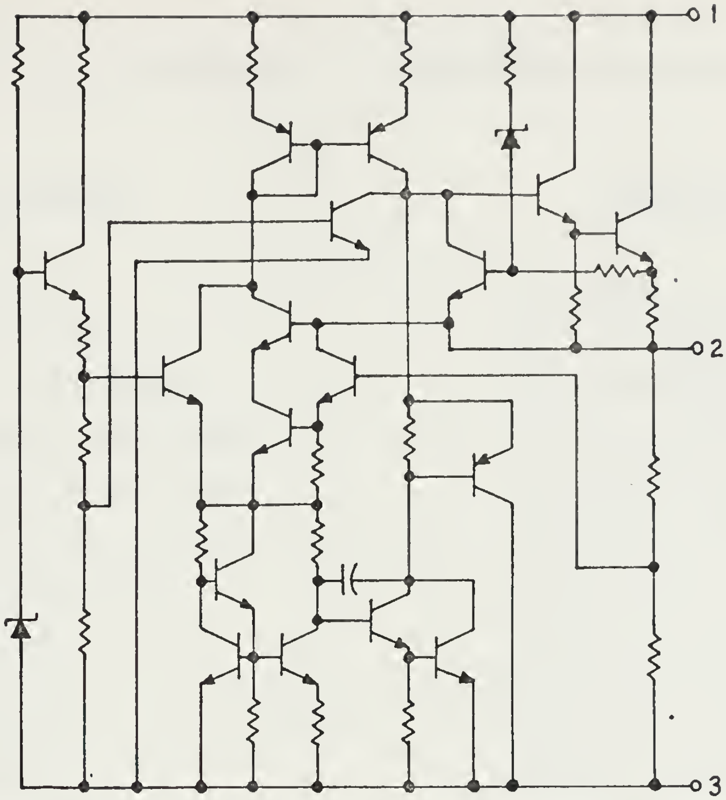

FIG. 58: MOTOROLA MC7800 SERIES

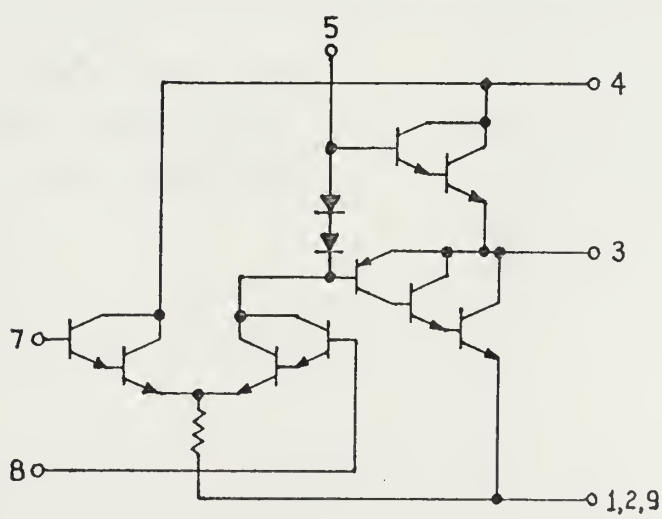

FIG. 59: MOTOROLA MFC9020

APPENDIX E   TEST EQUIPMENT LIST

Instrument:	Brand/Type:
Oscilloscope	Tektronics 531A
RF Signal Generator	USN SG-85/URM-25D
AF Signal Generator	USN AN/URM-127
Vacuum Tube Volt Meter	Hewlett-packard 400C
Vacuum Tube Volt Meter	Hewlett-packard 410B
Counter-Timer	Monsanto 100B
Digital Multimeter	Weston 1240
Power Supply	Hewlett-Packard 721A
Power Supply	Hewlett-Packard 6215A
Power Supply	Hewlett-Packard 6215A
Power Supply	Hewlett-Packard 6216A
Power Supply	Hewlett-Packard 6218A
Impedance Bridge	General Radio 1650-A
Q Meter	Boonton 260-A
Oscilloscope Camera	Polaroid CR-9
Decade Resistor	General Radio K134B
Capacitor Substitution Box	Eico 1120
Curve Tracer Oscilloscope	Tektronics 576
Sine-Square Wave Generator	EE-2216 Lab Project
IF Sweeper	Ligna-Sweep SKV 935C
Tube Tester	USN TV-10B/U

LIST OF REFERENCES

1. Knietel, Tom, ed., Surplus Conversion Handbook, p. 157-165, Cowan Publishing Co., 1964.

2. Blakeslee, Douglas, ed., The Radio Amateur's Handbook 1972, 47th ed., American Radio Relay League, 1972.

3. Technical Manual TM11-866 Radio Receivers BC-779-A,B; etc., Department of the Army, 1948.

4. The Plessey Co., Plessey Semiconductors SL600 Series Applications Manual.

5. Eslick, Bill, "The Ham Notebook," Ham Radio, v. 2, p. 68-69, November 1969.

6. Schultz, John J., "Integrated Circuit Noise Blanker," Ham Radio, v. 2, p. 52-55, May 1969.

7. Cross, Hank, "Converting a Vacuum-Tube Receiver to Solid State," Ham Radio, v. 2, p. 26-36, February 1969.

8. National Semiconductor Corp., Linear Integrated Circuits, 1973.

9. Hejhall, Roy, "An Integrated Circuit Balanced Modulator," Ham Radio, v. 3, p. 6-13, September 1970.

10. Motorola Semiconductor Products, Inc., _Linear Integrated Circuits Data Book_, 3rd ed., 1964.

11. Hrivnak, Paul, "Integrated Circuit Communications Receiver for 80 Meters," _Ham Radio_, v. 4, p. 6-13, July 1971.

12. Megirian, Ray, "A Complete Audio Module," _Ham Radio_, v. 6, p. 18-22, June 1973.

13. Math, Irwin, "Math's Notes," _CQ_, v. 29, p. 48-49, April 1973.

14. Trulove, James E., "Three Terminal Voltage-Regulator ICs," _Ham Radio_, v. 6, p. 26-30, December 1973.

15. Ashe, Jim, "Grounding and Wiring," _Ham Radio_, v. 2, p. 44-47, June 1969.

16. Dahl, Kirk V., "New Symbology for Digital-Logic Diagrams," _QST_, v. 58, p. 17-21, August 1974.

17. International Telephone and Telegraph Corp., _Reference Data for Radio Engineers_, 5th ed., Howard W. Sams & Co., 1968.

18. Millman, J. and Halkias, C. C., _Integrated Electronics: Analog and Digital Circuits and Systems_, McGraw-Hill Book Co., 1972.

19. Motorola Semiconductor Products, Inc., _The Integrated Circuit Data Book_, 1st ed., 1968.

20. Motorola Semiconductor Products, Inc., *The Semiconductor Data Library*, Reference Volume, 4th ed., 1973.

21. Motorola Semiconductor Products, Inc., *The Semiconductor Data Library*, Vol. I, 1st ed., 1972.

22. Motorola Semiconductor Products, Inc., *The Semiconductor Data Library*, Vol. II, 2nd ed., 1973.

23. Olson, Hank, "The Crystal Oscillator," *Ham Radio*, v. 2, p. 33-43, July 1969.

24. Orr, William I., *Radio Handbook*, 19th ed., Howard W. Sams & Co., 1972.

25. RCA Corp., *RCA Linear Integrated Circuits and MOS Devices*, 1973.

26. RCA Corp., *RCA Receiving Tube Manual*, 1971.

27. Signetics Corp., *Signetics Integrated Circuits*, 1972.

INITIAL DISTRIBUTION LIST

		No. Copies
1.	Defense Documentation Center Cameron Station Alexandria, Virginia 22314	2
2.	Library, Code 0212 Naval Postgraduate School Monterey, California 93940	2
3.	Professor Charles H. Rothauge, Code 52 Ro (thesis advisor) Department of Electrical Engineering Naval Postgraduate School Monterey, California 93940	1
3.	Professor W. Conley Smith, Code 52 Sm Department of Electrical Engineering Naval Postgraduate School Monterey, California 93940	1
5.	LT James M. Steussy, USN (student) Class 1-75 Naval Destroyer School Newport, Rhode Island 02840	1
6.	LT George L. Mager, USN Naval Security Group Activity Fort George G. Meade, Maryland 20755	1

7. LT Clifford J. Appel, USCG 1
CCGD 14 (EEE)
677 Ala Moana Blvd.
Honolulu, Hawaii 96813

8. LT Dan L. Wojtkowiak, USN 1
SMC 1454
Naval Postgraduate School
Monterey, California 93940

9. Professor Clarence F. Klamm, Jr., Code 52Kl 1
Department of Electrical Engineering
Naval Postgraduate School
Monterey, California 93940

Thesis 155505
S7125 Steussy
c.1 Solid-state conversion
of a tube-type communi-
cations receiver.